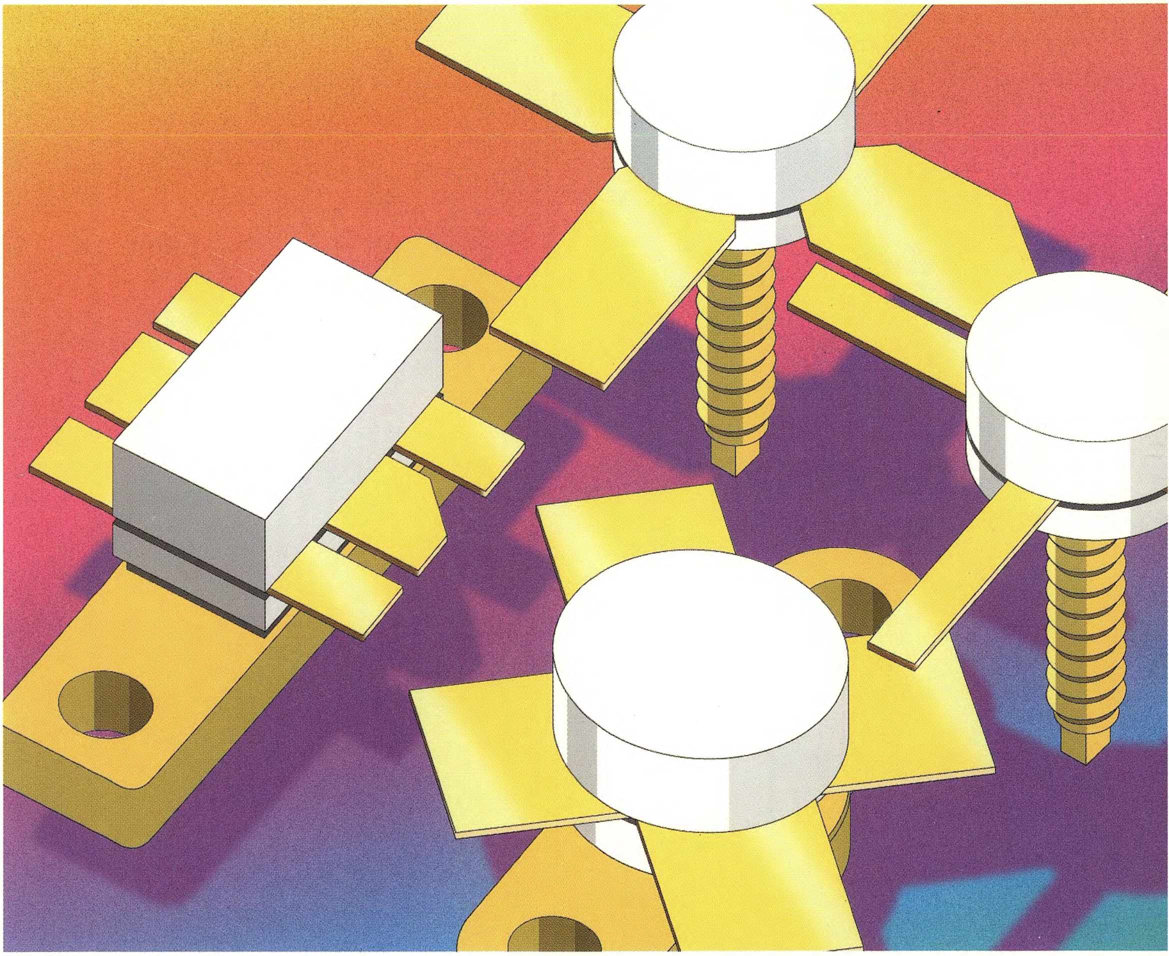
DISCRETE SEMICONDUCTORS

Mos & Bipolar Transmitting Transistor



1996

APPLICATION REPORTS

Philips Semiconductors

PHILIPS

RF POWER TRANSISTORS FROM PHILIPS SEMICONDUCTORS

Dear Customer,

Our Data Handbook for RF power transistors is currently undergoing a major revision to reflect the recent changes in our product range. We expect the new edition of our handbook to be available shortly but in the meantime, we are reprinting the earlier version. To assist you in using it, we indicate here the changes, i.e. deletions and replacements in our range.

Products omitted from range	Replacement	Products omitted from range	Replacement
BLT90/SL	–	BLX91A	BLW80
BLU10/12	–	BLX91CB	–
BLU15/12	–	BLX92A	BLW89
BLU30/28	–	BLX94A	BLX94C
BLU97/SL	BLU97	BLX95	–
BLV11	–	BLY87A	BLY87C/01
BLV37	–	BLY88A	BLY88C/01
BLV38	–	BLY90	–
BLV90/SL	BLV90	BLY91A	BLY91C/01
BLV91	BLV91/SL	BLY92A	BLY92/01
BLV94	BLV194	2N3375	–
BLV97	BLV97CE	2N3632	–
BLV98	BLV98CE	2N3852	–
BLW60	BLW60C	2N3924	–
BLW99	–	2N3926	–
BLX67	–	2N3927	–
BLX69A	–		

If you have any questions or comments concerning these or any products in our range, please don't hesitate to contact us at the address below.

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U.H.F. Transmitters

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CURRENT RANGE OF PHILIPS COMPONENT RF POWER TRANSISTORS

type	envelope	mode of operation	V_{CE} V	frequency MHz	output power W	power gain dB
BFQ42	TO-39/1	CW; class-B	13.5	175	2	> 11
			12.5	175	2	typ. 10.5
BFQ43	TO-39/3	CW; class-B	13.5	175	4	> 12
			12.5	175	4	typ. 12
BFQ43S	TO-39/3	CW; class-B	13.5	175	4	> 12
			12.5	175	4	typ. 12
BFS22A	TO-39/1	CW; class-B	13.5	175	4	> 8
			12.5	175	4	typ. 8
BFS23A	TO-39/1	CW; class-B	28	175	4	> 10
BLF ...	see MOSFETs page					
BLT90/SL	SOT172	CW; class-B	7.5	900	0.75	> 7
BLT91/SL	SOT172	CW; class-B	7.5	900	1.5	> 6
BLT92/SL	SOT122	CW; class-B	7.5	900	3	> 7
BLT93/SL	SOT122	CW; class-B	7.5	900	6	> 5.5
BLU20/12	SOT119	CW; class-B	12.5	470	20	> 6.5
BLU30/28	SOT119	CW; class-B	28	470	30 (note 5)	> 8
BLU30/28	SOT119	CW; class-B	24	470	25 (note 5)	typ. 8
BLU60/28	SOT119	CW; class-B	28	470	60 (note 5)	> 7
BLU60/28	SOT119	CW; class-B	24	470	50 (note 5)	typ. 7
BLU30/12	SOT119	CW; class-B	12.5	470	30	> 6
						typ. 7.4
BLU45/12	SOT119	CW; class-B	12.5	470	45	> 4.8
BLU50	SOT161	CW; class-B	28	400	30	> 10
BLU51	SOT161	CW; class-B	28	400	45	> 9
BLU52	SOT161	CW; class-B	28	400	60	> 8
BLU53	SOT161	CW; class-C	28	400	100	> 7
BLU60/12	SOT119	CW; class-B	12.5	470	60	> 4.4
BLU97	SOT122	CW; class-B	12.5	470	7	> 8.5
BLU98	SOT103	CW; class-B	12.5	900	0.5	> 8
BLU99	SOT122	CW; class-B	12.5	470	5	> 10.5
			12.5	900	4	typ. 7
BLV10	SOT123	CW; class-B	13.5	175	8	> 9
			12.5	175	8	typ. 10.5
		SSB; class-A	12	28	1 (note 3)	18
BLV11	SOT123	CW; class-B	13.5	175	15	> 8
			12.5	175	15	typ. 7.5
		SSB; class-A	12	28	2 (note 3)	18
		SSB; class-AB	13.5	28	10 (note 4)	18
BLV20	SOT123	CW; class-B	28	175	8	> 12
		SSB; class-A	26	28	1.3 (note 3)	20

Notes: see next page.

type	envelope	mode of operation	V_{CE} V	frequency MHz	output power W		power gain dB
BLV21	SOT123	CW; class-B	28	175	15		> 10
		SSB; class-A	26	28	2.3	(note 3)	20
BLV25	SOT119	CW; class-B narrow band	28	108	175		> 10
BLV30	SOT122	lin.ampl., class-A	25	225	1.5	(note 1)	> 18
			25	225	1.7	(note 1)	typ. 20
BLV31	SOT122	lin.ampl., class-A	25	225	5	(note 1)	> 15
			25	225	7	(note 1)	typ. 16.5
BLV32F	SOT160	lin.ampl., class-A	25	225	10	(note 2)	> 16
			25	225	12.5	(note 2)	typ. 17.2
BLV33	SOT147	lin.ampl., class-A	25	225	19	(note 2)	> 9
			25	225	26	(note 2)	typ. 9.7
		class-AB	28	225	90	(note 2)	typ. 6.5
BLV33F	SOT119	lin.ampl., class-A	25	225	16	(note 2)	> 13.5
			25	225	22	(note 2)	typ. 14.8
		class-AB	28	225	85	(note 2)	typ. 10.5
BLV36	SOT161	lin.ampl., class-AB	28	225	115		> 11
			28	225	115		typ. 13
			12.5	175	8		typ. 10.5
BLV37	SOT179	CW; class-B	28	108	250		> 10.5
							typ. 11.3
BLV38	SOT179	lin.ampl., class-AB	35	225	225		> 8
							typ. 8.8
BLV45/12	SOT119	CW; class-B	12.5	175	45		> 6.5
BLV57	SOT161	lin.ampl., class-A	25	860	6	(note 2)	> 8
			25	860	12	(note 2)	typ. 9
		lin.ampl., class-AB	25	860	38		typ. 6.5
BLV59	SOT171	lin.ampl., class-AB	25	860	30		7
BLV75/12	SOT119	CW; class-B	12.5	175	75		> 6.5
							typ. 7.5
BLV80/28	SOT121	CW; class-B	28	175	80		> 6.5
							typ. 7
BLV90	SOT172	CW; class-B	12.5	900	1		> 7.5
			9.6	900	0.75		typ. 7.9
BLV90/SL	SOT172	CW; class-B	12.5	900	1		> 7.5
			9.6	900	1		typ. 7
BLV91	SOT172	CW; class-B	12.5	900	2		> 6.5
			9.6	900	1.5		typ. 6.6

Notes

1. $P_{o\ sync}$ at $d_{im} < -60$ dB.
2. $P_{o\ sync}$ at $d_{im} < -55$ dB.
3. PEP at $d_3 < -40$ dB.
4. PEP at d_3 typ. -30 dB.
5. Available on request as loose-leaf data

type	envelope	mode of operation	V_{CE} V	frequency MHz	output power W	power gain dB
BLV91/SL	SOT172	CW; class-B	12.5 9.6	900 900	2 1.5	> 6.5 typ. 6.6
BLV92	SOT171	CW; class-B	12.5 9.6	900 900	4 3	> 7.5 typ. 7.3
BLV93	SOT171	CW; class-B	12.5 9.6	900 900	8 6	> 6.5 typ. 6
BLV94	SOT171	CW; class-B	12.5	900	15	> 6 typ. 7
BLV95	SOT171	CW; class-B	12.5	900	22	> 5.5
BLV97	SOT171	CW; class-B	24	900	30	> 7 typ. 8
BLV98	SOT171	CW; class-B	24	900	14	> 8.5 typ. 10
BLV99	SOT172	CW; class-B	24	900	2	> 8 typ. 9.3
BLW29	SOT120	CW; class-B	13.5 12.5	175 175	15 15	> 10 typ. 10.5
BLW31	SOT120	CW; class-B	13.5 12.5	175 175	28 28	> 9 typ. 9.5
BLW32	SOT122	lin.ampl., class-A	25 25	860 860	0.5 (note 1) 0.63 (note 1)	> 11 typ. 12.2
BLW33	SOT122	lin.ampl., class-A	25 25	860 860	1 (note 1) 1.15 (note 1)	> 10 typ. 10.5
BLW34	SOT122	lin.ampl., class-A	25 25	860 860	1.8 (note 1) 2.15 (note 1)	> 9 typ. 10.2
BLW50F	SOT123	SSB; class-A SSB; class-AB	45 50	1.6-28 1.6-28	0-16 (note 3) 10-65 (note 4)	> 19.5 typ. 18
BLW60	SOT56	CW; class-B SSB; class-AB	12.5 12.5	175 1.6-28	45 3-30 (note 4)	> 5 typ. 19.5
BLW60C	SOT120	CW; class-B SSB; class-AB	12.5 12.5	175 1.6-28	45 3-30 (note 4)	> 5 typ. 19.5
BLW76	SOT121	SSB; class-AB CW; class-B	28 28	1.6-28 108	8-80 (note 4) 80	> 13 typ. 7.9
BLW77	SOT121	SSB; class-AB CW; class-B	28 28	1.6-28 87.5	15-130 (note 4) 130	> 12 typ. 7.5
BLW78	SOT121	CW; class-B SSB; class-A SSB; class-AB	28 26 28	150 28 28	100 35 (note 3) 100 (note 4)	> 6 typ. 19.5 typ. 19

Notes

1. $P_{o\ sync}$ at $d_{im} < -60$ dB.
2. $P_{o\ sync}$ at $d_{im} < -55$ dB.
3. PEP at $d_3 < -40$ dB.
4. PEP at d_3 typ. -30 dB.

type	envelope	mode of operation	V_{CE} V	frequency MHz	output power W	power gain dB
BLW79	SOT122	CW; class-B	12.5	470	2	> 9
			12.5	175	2	typ. 13.5
BLW80	SOT122	CW; class-B	12.5	470	4	> 8
			12.5	175	4	typ. 15
BLW81	SOT122	CW; class-B	12.5	470	10	> 6
			12.5	175	10	typ. 13.5
BLW83	SOT123	SSB; class-A	26	1.6-28	0-10 (note 3)	> 20
		SSB; class-AB	28	1.6-28	3-30 (note 4)	typ. 21
BLW84	SOT123	CW; class-B	28	175	25	> 9
BLW85	SOT123	CW; class-B	12.5	175	45	> 4.5
			13.5	175	45	typ. 6
		SSB; class-AB	12.5	1.6-28	3-30 (note 4)	typ. 19.5
BLW86	SOT123	CW; class-B	28	175	45	> 7.5
		SSB; class-AB	28	1.6-28	5-47.5 (note 4)	typ. 19
		SSB; class-A	26	1.6-28	17 (note 3)	typ. 22
BLW87	SOT123	CW; class-B	13.5	175	25	> 6
						typ. 6.6
BLW89	SOT122	CW; class-B	28	470	2	> 12
						typ. 13.5
BLW90	SOT122	CW; class-B	28	470	4	> 11
						typ. 12.5
BLW91	SOT122	CW; class-B	28	470	10	> 9
						typ. 10.5
BLW95	SOT121	SSB; class-AB	50	1.6-28	20-160 (note 4)	> 14
BLW96	SOT121	SSB; class-AB	50	1.6-28	25-200 (note 4)	> 13.5
		CW; class-B	50	108	200	typ. 6.5
		SSB; class-A	40	28	50 (note 3)	typ. 19
BLW97	SOT121	SSB; class-AB	28	1.6-28	175 (note 4)	> 11.5
BLW98	SOT122	lin.ampl., class-A	25	860	3.5 (note 1)	> 6.5
			25	860	4.4 (note 1)	typ. 7
BLW99	SOT121	SSB; class-AB	12.5	1.6-28	80 (note 4)	> 12.5
BLX13	SOT56	SSB; class-A	26	28	0-8 (note 3)	> 18
		SSB; class-AB	28	28	25 (note 4)	> 18
		CW; class-B	28	70	25	typ. 17
BLX13C	SOT120	SSB; class-A	26	1.6-28	0-8 (note 3)	> 20
		SSB; class-AB	28	1.6-28	3-25 (note 4)	typ. 21
BLX14	SOT55	SSB; class-A	28	1.6-28	25 (note 3)	> 13
		SSB; class-AB	28	1.6-28	7.5-50 (note 4)	> 13
		CW; class-B	28	70	50	> 7.5
		CW; class-B	28	30	50	typ. 16

Notes

1. $P_{o\ sync}$ at $d_{im} < -60$ dB.
2. $P_{o\ sync}$ at $d_{im} < -55$ dB.
3. PEP at $d_3 < -40$ dB.
4. PEP at d_3 typ. -30 dB.

type	envelope	mode of operation	V_{CE} V	frequency MHz	output power W	power gain dB
BLX15	SOT55	SSB; class-AB	50	1.6-28	20-150 (note 4)	> 14
		SSB; class-A	40	1.6-28	30 (note 3)	> 14
		CW; class-B	50	70	150	> 10
		CW; class-B	50	108	150	typ. 7.4
BLX39	SOT120	CW; class-B	28	175	45	> 7.5
		SSB; class-AB	28	1.6-28	5-42.5 (note 4)	typ. 19
		SSB; class-A	26	1.6-28	15 (note 3)	typ. 20
BLX65	TO-39/1	CW; class-B	13.8	470	2	typ. 7
		CW; class-B	12.5	470	2	> 6
		CW; class-B	12.5	175	2	typ. 12
BLX65E	TO-39/3	CW; class-B	12.5	175	2	typ. 16
		CW; class-B	12.5	470	2	> 9
BLX65ES	TO-39/3	CW; class-B	12.5	175	2	typ. 16
		CW; class-B	12.5	470	2	> 9
BLX67	SOT48/3	CW; class-B	13.8	470	1.5	typ. 10
		CW; class-B	13.8	470	3.0	typ. 9.3
			12.5	470	2.5	> 8.5
			12.5	175	3.0	typ. 20
BLX68	SOT48/3	CW; class-B	13.8	470	7	> 5.4
			13.8	470	7.8	typ. 5.9
			12.5	470	7	> 5
			12.5	175	7.2	typ. 12.6
BLX69A	SOT48/2	CW; class-B	13.5	470	20	> 4
			12.5	470	17	> 4
			12.5	175	17	typ. 11
BLX91A	SOT48/3	CW; class-B	24	470	0.85	typ. 12.3
			28	470	1	> 11
			28	470	1.45	typ. 12.6
			28	1000	1.4	typ. 5.4
BLX91CB	SOT48/3	video CRT driver	28	"V_{CESM} max. 65 V; C_c typ. 3 pF"		
BLX92A	SOT48/3	CW; class-B	24	470	2.4	typ. 10.8
			28	470	2.5	> 11
			28	470	3	typ. 11.7
			28	1000	2.5	typ. 5.5
BLX93A	SOT48/3	CW; class-B	24	470	7	typ. 8.5
			28	470	7	> 8.5
			28	470	8	typ. 9
			28	1000	5	typ. 5.2
BLX94A	SOT48/2	CW; class-B	28	470	25	> 6
						typ. 6.5
BLX94C	SOT122	CW; class-B	28	470	25	> 6 5
						typ. 7.25

Notes

1. $P_{o\ sync}$ at $d_{im} < -60$ dB.
2. $P_{o\ sync}$ at $d_{im} < -55$ dB.
3. PEP at $d_3 < -40$ dB.
4. PEP at d_3 typ. -30 dB.

type	envelope	mode of operation	V_{CE} V	frequency MHz	output power W	power gain dB
BLX95	SOT56	CW; class-B	28 28	470 175	40 40	< 4.5 typ. 11
BLX96	SOT48/3	lin.ampl., class-A	25 25	860 860	0.5 (note 1) 0.6 (note 1)	> 6 typ. 7
BLX97	SOT48/3	lin.ampl., class-A	25 25	860 860	1 (note 1) 1.1 (note 1)	> 5.5 typ. 6.5
BLX98	SOT48/2	lin.ampl., class-A	25 25	860 860	3.5 (note 1) 4.0 (note 1)	> 5 typ. 5.5
BLY87A	SOT48/2	CW; class-B CW; class-B	13.5 12.5	175 175	8 8	> 9 typ. 9
BLY87C	SOT120	CW; class-B CW; class-B	13.5 12.5	175 175	8 8	> 12 typ. 11.5
BLY88A	SOT48/2	CW; class-B CW; class-B	13.5 12.5	175 175	15 15	> 7.5 typ. 7.5
BLY88C	SOT120	CW; class-B CW; class-B	13.5 12.5	175 175	15 15	> 8 typ. 7.5
BLY89A	SOT56	CW; class-B	13.5	175	25	> 6
BLY89C	SOT120	CW; class-B	13.5	175	25	> 6 typ. 6.6
BLY90	SOT55	CW; class-B	12.5	175	50	> 5
BLY91A	SOT48/2	CW; class-B	28	175	8	> 12
BLY91C	SOT120	CW; class-B	28	175	8	> 12
BLY92A	SOT48/2	CW; class-B	28	175	15	> 10
BLY92C	SOT120	CW; class-B	28	175	15	> 10
BLY93A	SOT56	CW; class-B	28	175	25	> 9
BLY93C	SOT120	CW; class-B	28	175	25	> 9
BLY94	SOT55	CW; class-B	28	175	50	> 7
2N3375	TO-60	CW; class-B CW; class-B	28 28	100 400	7.5 > 3	> 8.8 > 4.8
2N3553	TO-39/1	CW; class-B	28	175	2.5	> 10
2N3632	TO-60	CW; class-B	28	175	> 13.5	> 5.9
2N3866	TO-39/1	CW; class-B	28	400	1	> 10
2N3924	TO-39/1	CW; class-B	13.5	175	4	> 6
2N3926	TO-60	CW; class-B	13.5	175	7	> 5.4
2N3927	TO-60	CW; class-B	13.5	175	12	> 4.8
2N4427	TO-39/1	CW; class-B	12	175	1	> 10

Notes

1. $P_{o\ sync}$ at $d_{im} < -60$ dB.
2. $P_{o\ sync}$ at $d_{im} < -55$ dB.
3. PEP at $d_3 < -40$ dB.
4. PEP at d_3 typ. -30 dB.

type (MOSFETs)	envelope	mode of operation	V_{DS} V	frequency MHz	output power W		power gain dB
BLF145	SOT123	SSB; class-A	28	1.6-28	8	(note 3)	> 24
		SSB; class-AB	28	1.6-28	30	(note 4)	typ. 20
BLF147	SOT121	SSB; class-AB	28	1.6-28	150	(note 4)	> 17
BLF175	SOT123	SSB; class-A	50	1.6-28	8	(note 3)	> 24
		SSB; class-AB	50	1.6-28	30	(note 4)	typ. 23
BLF177	SOT121	SSB; class-AB	50	1.6-28	150	(note 4)	> 20
		CW; class-B	50	108	150		typ. 19
BLF221	TO-39/3	CW; class-B	12.5	175	2	(note 5)	> 10
BLF241	SOT5/11	CW; class-AB	12.5	175	2		> 10
		CW; class-B	28	175	3		typ. 14
BLF242	SOT123	CW; class-AB	28	175	5		> 13
		CW; class-B	28	400	5		13
BLF244	SOT123	CW; class-B	28	175	15		> 13
		SSB; class-A	28	28	4	(note 3)	typ. 24
		CW; class-B	28	400	15		typ. 11
BLF245	SOT123	CW; class-B	28	175	30		> 13
		CW; class-B	28	400	30		typ. 10
BLF246	SOT121	CW; class-B	28	108	80		> 16
		CW; class-B	28	28	80		typ. 20
BLF278	NO-298	CW; class-B	50	108	300	(note 5)	> 20
BLF368	NO-298	lin.ampl., class-AB	32	225	300	(note 5)	> 12
BLF378	NO-298	lin.ampl., class-AB	50	225	250	(note 5)	> 13
BLF521	SOT172D	CW; class-B	12.5	500	2	(note 5)	> 10
BLF522	SOT171	CW; class-B	12.5	500	5	(note 5)	> 10
BLF543	SOT171	CW; class-B	28	500	10	(note 5)	> 12
BLF544	NO-297	CW; class-B	28	500	20	(note 5)	> 12
BLF545	NO-297	CW; class-B	28	500	40	(note 5)	> 11
BLF547	NO-294	CW; class-B	28	500	100	(note 5)	> 10
BLF548	NO-298	CW; class-B	28	500	150	(note 5)	> 10

Notes

3. PEP at $d_3 < -40$ dB.
4. PEP at d_3 typ. -30 dB.
5. Available on request as loose-leaf data.

PHILIPS

Application information

Electronic components and materials

AI 534

A survey of developments in R. F. power amplifiers up to 300 W P.E.P. output

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A Survey of Developments in R.F. Power Amplifiers up to 300 W P.E.P. Output

Date of release: 5 October 1973

Note

For a general explanation of the symbols referring to transmitting transistors, see Data Handbook Part SC4a. Some symbols and abbreviations appearing in this publication are as follows:

B	bandwidth
d_3, d_5	two-tone i.m.d. 3rd and 5th order
eff'y	efficiency
f_1, f_2	frequencies used in 2-tone i.m.d. measurement
G_p	power gain
I_{CZS}	collector current at zero signal
i.m.d.	intermodulation distortion
I_T	total current from supply
P_i or P_S	input or source power
P_o or P_L	output or load power
P_{refl}	reflected power from amplifier under test
O'all eff'y	overall efficiency
T_h	heatsink temperature
V_B or V_{CC}	supply voltage
V_{CE}	collector-emitter voltage
η_{dt}	double-tone efficiency
$W_{peak\ sync}$	peak envelope power over sync pulses in a TV vision signal

INTRODUCTION

In the communications field there is a continuing need for solid-state r.f. power amplifiers giving outputs up to 300 W P.E.P. Such amplifiers may form the final stages of low power mobile transmitters (e.g. fire, police, ambulance, amateur vehicle systems; marine ship-to-shore installations) or they may be used as driver stages in larger radio/TV broadcast transmitters. Our range of transmitting transistors includes devices suitable for all these applications, and it is the purpose of this publication to bring to general notice the circuit development work that has been done in our laboratories in this field.

The survey is divided into three parts, i.e. TV applications, general VHF/UHF applications and S.S.B. applications. In each part, circuit diagrams and performance data for various amplifiers are given. Copies of any Application Report mentioned in this survey can be obtained by quoting the appropriate reference. Full information on the semiconductors is available in the Data Handbook (Part SC4a).

TELEVISION AMPLIFIERS

BLY93A AS A LINEAR TV AMPLIFIER

In TV transposers, where vision and sound signals are amplified together, the linearity requirements are so stringent that they can be met only with a Class-A amplifier. TV transmitters, on the other hand, amplify vision and sound signals separately and so have a lower linearity requirement; this can be met with a Class-AB amplifier thus obtaining higher output power and efficiency. Power amplifiers have been developed using the BLY93A for operation in Class-A at Band I frequencies (Fig.1), and in Class-A and -AB at Band III frequencies (Fig.2).

As a driver stage in TV transmitters the amplifier can deliver an output power of 12 $W_{peak\ sync}$ under the following conditions:

sync compression	26/25 %
differential gain	95 %
differential phase	1°

The amplifier can also be used in TV transposers. As a final stage, it can deliver 10 $W_{peak\ sync}$ at a 3-tone i.m.d. level of −52 dB or, for driver applications, this becomes 4,5 $W_{peak\ sync}$ at an i.m.d. of −60 dB.

This report has two appendices:

Appendix A – Correlation Between Two-tone and Three-tone Tests

Appendix B – TV System Characteristics and TV Transmitter and Transposer Requirements.

The second appendix gives some information on:

(*a*) TV system characteristics as recommended by C.C.I.R.

(*b*) Transmitter and transposer systems in common use

(*c*) Description and definition of transmitter and transposer requirements concerning:
   (i) Frequency response
   (ii) Amplitude and phase non-linearity
   (iii) Intermodulation distortion.

Band I Amplifier

The amplifier described[1]) and shown in Fig.1 is tuned to Channel $E3$ (54-61 MHz). The adjustment of the BLY93A is:

$V_{CE} = 28$ V $\qquad I_C = 1$ A.

Some important properties are:

Power gain	19 dB
Bandwidth at -1 dB	15 MHz
Input VSWR	<1,1

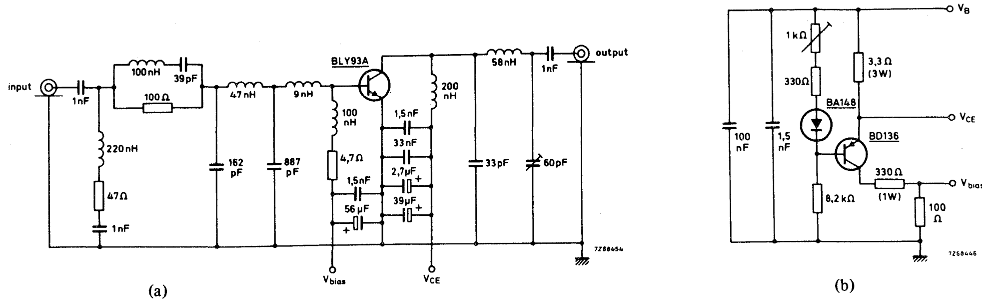

Fig.1. (a) Band I amplifier and (b) its Class-A bias circuit.

Band III Amplifier

The amplifier described[2]) and shown in Fig.2 can be tuned to any channel in Band III (174-230 MHz); it can be biased either in Class-A or in Class-AB.

When biased in Class-A, the adjustment of the transistor is

$V_{CE} = 28$ V $\qquad I_C = 1$ A.

Amplifier properties are:

Power gain	11,75 dB (tuned to channel E_{12})
Bandwidth at −1 dB	30 MHz
Input VSWR	<1,1 over 6 MHz.

As a driver stage in TV transmitters, the amplifier can deliver an output power of 12 $W_{peak\ sync}$ under the following conditions:

sync. compression	27/25 %
differential gain	98 %
differential phase	<1°

The amplifier can also be used in TV transposers. As a final stage it can deliver 8,25 $W_{peak\ sync}$ at a 3-tone i.m.d. level of -52 dB or, for driver applications, this becomes 4,75 $W_{peak\ sync}$ at an i.m.d. of -60 dB.

When this amplifier is biased in Class-AB, the transistor adjustment is:

$V_{CE} = 28$ V $\qquad I_{CZS} \approx 40$ mA.

The amplifier properties in this case are:

Power gain	9,25 dB (tuned to channel E_{12})
Bandwidth at -1 dB	19 MHz
Input VSWR	<1,1 over 5 MHz.

When used as a driver stage in TV transmitters, the amplifier can deliver an output power of 16 $W_{peak\ sync}$ under the following conditions:

sync compression	27/25 %
differential gain	98 %
differential phase	3°

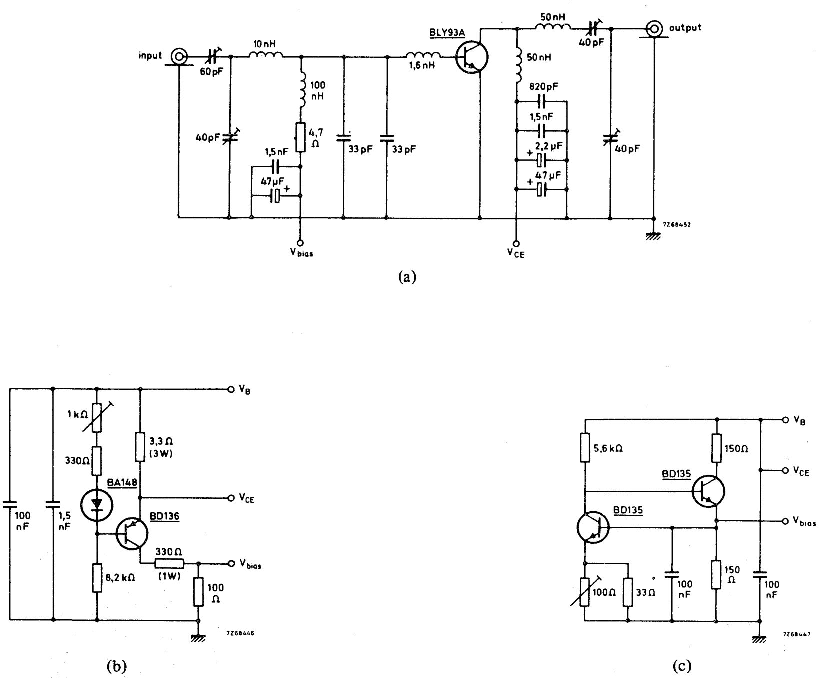

Fig.2. (a) Band III amplifier; (b) class-A bias circuit and (c) class-AB bias circuit.

VHF/UHF APPLICATIONS

FREQUENCY TRIPLER (157-470 MHz) USING BFR63[3])

This is a 12 V circuit suitable for use in mobile transmitters in the 470 MHz band. The circuit (Fig.3) gives an output power of at least 400 mW, with a minimum gain of 6 dB and efficiency better than 30 %. Source and load impedances are both 50 Ω.

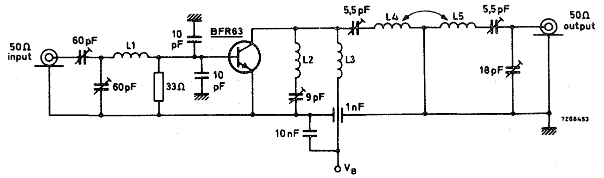

Fig.3.

ELECTRONIC ANTENNA SWITCHES FOR THE 160 MHz BAND USING BA182 DIODES[4])

This is a wide-band untuned electronic antenna switch covering the frequency range 132-174 MHz. The circuit includes two BA182 and one BAX13 diodes. The switching system is analogous to the use of T-R cells in radar systems. In this circuit, the λ/4 line sections are replaced by the equivalent lumped circuits, and the diodes take the place of the T-R cells. Though designed to handle power levels of 12 W, tests have been made at double this power to ensure safe operation at the lower power level. The insertion loss is approximately 0,5 dB in both transmit and receive conditions; receiver isolation during transmission is approximately 26 dB.

Two circuits are discussed, the first of which (Fig.4) is unbiased and therefore fully automatic. However, since this can be used only with a constant carrier type of transmitter, it is not suitable for amplitude modulation systems. A second circuit (Fig.5) is given which is suitable for AM transmitters but needs more components and an external voltage.

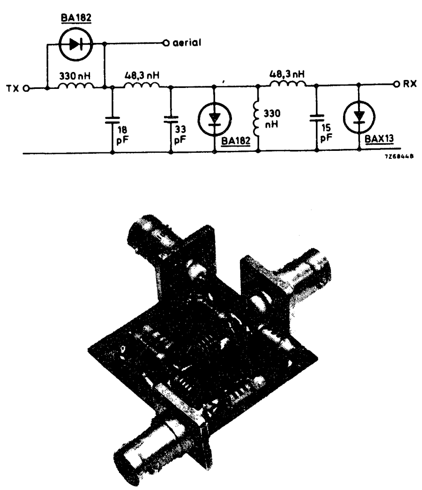

Fig.4.

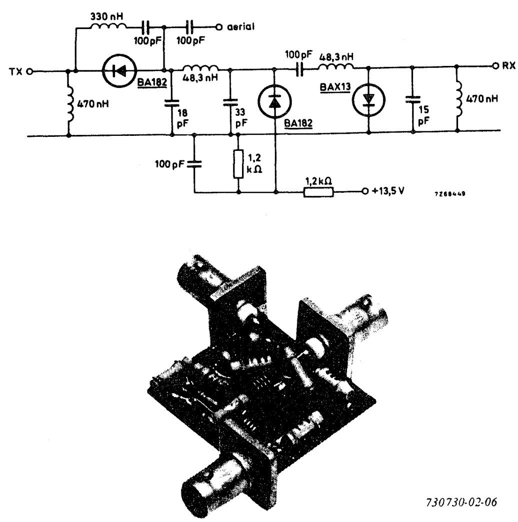

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Fig.5.

AMPLIFIERS FOR THE 470 MHz BAND

Three-stage 15 W Power Amplifier[5])

This amplifier (Fig.6) is suitable for mobile R/T transmitters with an aerial power of 15 W CW, tunable over the frequency range 430-470 MHz and operating from a 12 V supply. The three stages are: pre-driver BLX67, driver BLX68, final stage BLX69. The amplifier has the following characteristics:

f = 470 MHz. Load and source impedance = 50 Ω.

V_B (V)	P_i (mW)	P_o (W)	Gain (dB)	I_T (A)	O'all eff'y (%)	B at -1 dB
12,5	166	15	18,55	2,68	45	21 MHz
13,8	106	15	21,5	2,41	45	15,5 MHz

V_B = 12,5 V $\qquad$ P_o = 15 W

f (MHz)	P_i (mW)	Gain (dB)	I_T (A)	O'all eff'y (%)
450	152	19,95	2,60	46
460	160	19,72	2,65	45,5
470	166	18,55	2,68	45

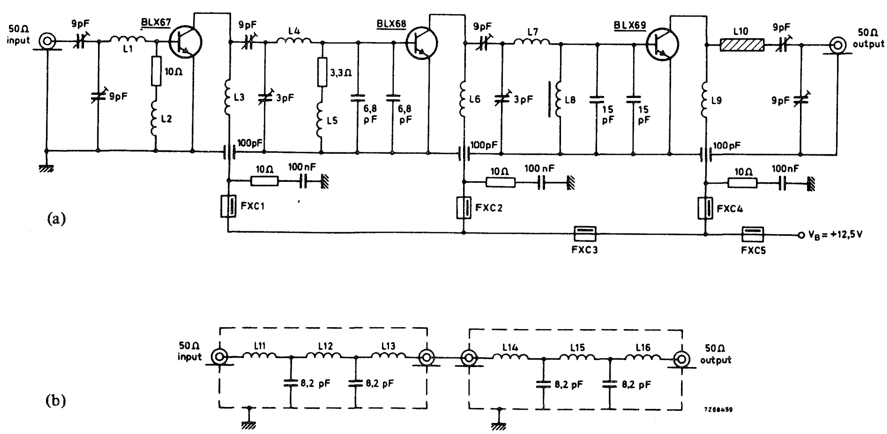


Fig.6. (a) 15 W 470 MHz power amplifier and (b) low-pass aerial filter giving 75 dB attenuation of second harmonic.

A Single-stage 40 W Power Amplifier[6])

The circuit of Fig.7 shows an amplifier for mobile R/T, with the somewhat high antenna power of 40 W from a 13,5 V supply, or 34 W from a 12,5 V supply. The amplifier uses two BLX69's connected in parallel. No antenna filter is included but, if desired, a filter such as that described in Reference 5 could be incorporated. Strip transmission lines are used at input and output.

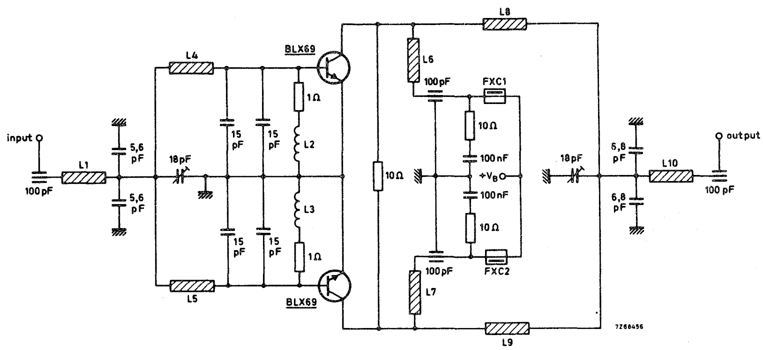
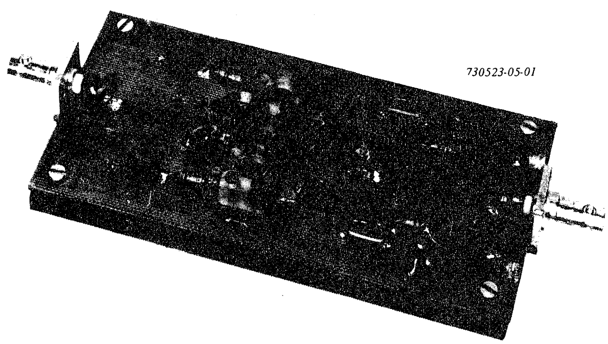


Fig.7.



Typical performance at 470 MHz:

V_B (V)	P_i (W)	P_o (W)	P_{refl} (W)	VSWR	Gain (dB)	I_T (A)	Eff'y (%)
13,5	13,5	40	0,02	1,06	4,72	4,3	69
12,5	12,0	34	0,025	1,1	4,53	4,0	68

This amplifier and the previous one (Ref. 5) are free of parasitic oscillations for an output VSWR up to 10, with variable phase (0-360°).

AMPLIFIERS FOR THE 160 MHz BAND

Two-stage Power Amplifier for 100 W at 175 MHz Using 2 x BLY90[7])

This amplifier uses a BLY90 in the driver stage and two parallel-connected BLY90's in the final stage. The circuit (Fig.8) has been calculated according to the design theory given in AI531 "Solid-state Power Amplifiers for the 160 MHz Band Using Emitter-grid Transistors".*

Typical performance at f = 175 MHz, T_h = 25°C
R_S and R_L = 50 Ω.

	P_i = 10 W	P_i = 15 W
V_B (V)	P_o (W)	P_o (W)
12,5	103	107
13,0	108	120
13,5	113	127

P_o measured 1 min after switch-on.

Two-stage Wide-band Amplifier with 12 W min Output[8])

This amplifier covers the frequency band 132-174 MHz with a gain variation over the band smaller than 0,5 dB. The circuit (Fig.9) uses a BLY87 in the driver stage and a BLY88 in the final stage.

Main properties of the amplifier are:

frequency range	132-174 MHz
supply voltage	13,8 V
output power	13,5 ± 1,5 W
input power	250 mW
reflected drive power	⩽ 5 mW (⩽2 %)
overall efficiency	⩾ 50 %

The matching networks are composed of Chebychev low-pass sections. In this report, special attention is given to the alignment of the amplifier and the swept-frequency measuring equipment. Design information and circuit diagram are given for a constant-gain amplifier (used in the test set-up) for 1 mW input, 250 mW output and gain variation of less than 0,1 dB over the stated frequency band.

THE BEHAVIOUR OF THE BLY91A AT LOW POWER LEVELS AND FREQUENCIES[9])

In designing r.f. power amplifiers it is sometimes expedient to use a transistor that is larger than necessary for the application in hand. However, under certain conditions, this can lead to instability in one or more stages. This report discusses the measures required to suppress such parasitic oscillations and takes the BLY91A as a test case. Though capable of an output of 8 W, its performance at the following combinations of frequency and power are given:

(i)	175 MHz	2 W
(ii)	175 MHz	1 W
(iii)	80 MHz	2 W
(iv)	80 MHz	1 W

* Order code. 9399 224 53101.

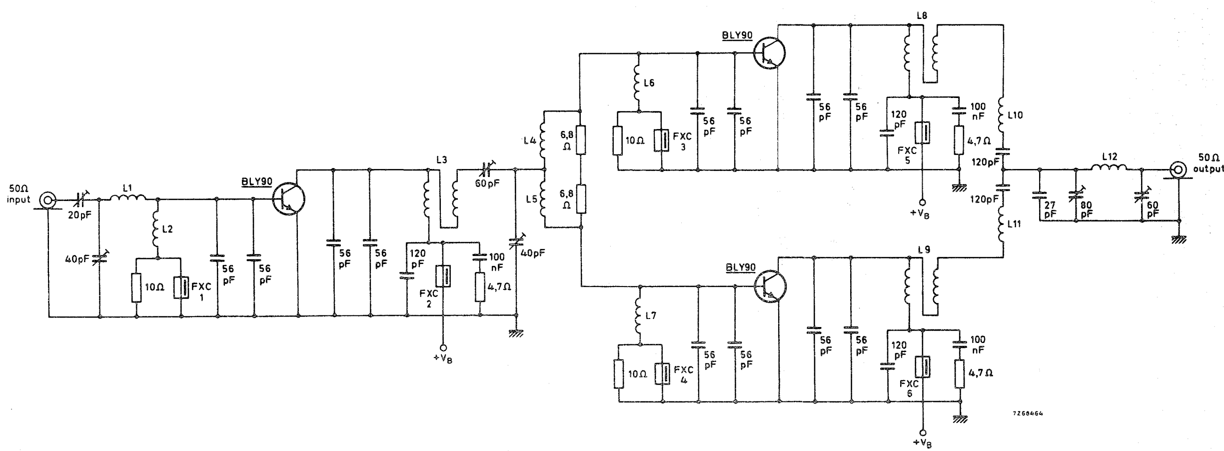


Fig.8. Power amplifier for 100 W at 175 MHz.

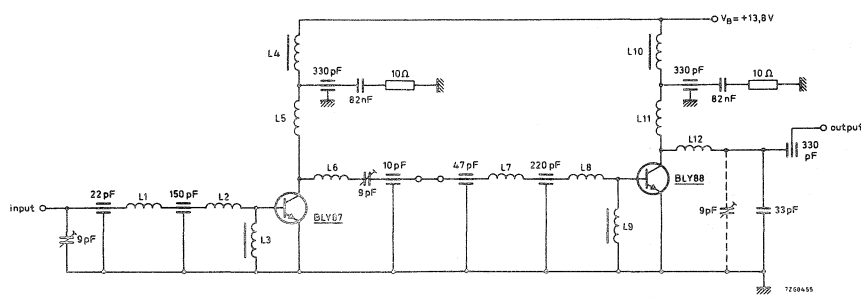


Fig.9. Wide-band amplifier (132-174 MHz) for 12 W output.

SINGLE-SIDEBAND APPLICATIONS

A 10 MHz 25 W P.E.P. LINEAR POWER AMPLIFIER USING BDY92 AT V_B = 20 V[10])

An existing amplifier[11]) has been modified to produce a linear power amplifier for use in s.s.b. systems. The modification consists of the addition of a two-transistor bias circuit; Fig.10 shows the circuit of the complete amplifier.

Typical performance at V_B = 20 V.

P_o (W PEP)	I_T (A)	2-tone eff'y %	d_3 (dB)	d_5 (dB)
20	1,4	35,5	–31	–56
25	1,6	39,2	–28	–56
30	1,76	42,5	–24	–55

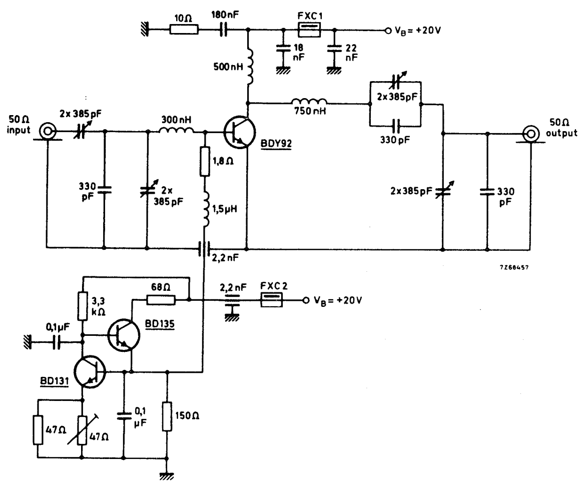


Fig.10.

WIDE-BAND AMPLIFIERS

All of the following linear power amplifiers cover the frequency range 1,6-28 MHz. The circuits include Class-A amplifiers for use as drivers, and Class-AB amplifiers for use as final stages.

Single-stage Driver Modules with BLY92A and BLX13 Operating in Class-A[12])

This report describes two separate amplifiers both of which are intended for driver applications in s.s.b. transmitters. The first amplifier(Fig.11) uses a BLY92A and delivers 3 W P.E.P. at an i.m.d. of better than –40 dB; its gain is approximately 18 dB over the band, and the input VSWR is less than 1,3. The second amplifier (Fig.12) uses a BLX13 and delivers 8 W P.E.P. at an i.m.d. of better than –40 dB; its gain is approximately 17 dB over the band, and the input VSWR is ≤ 1,5. Both amplifiers operate from a 28 V supply.

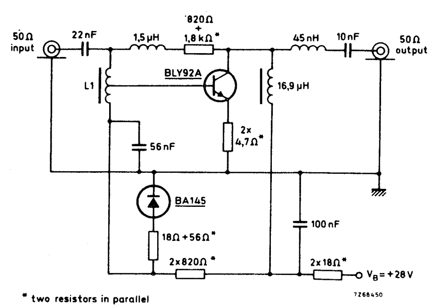
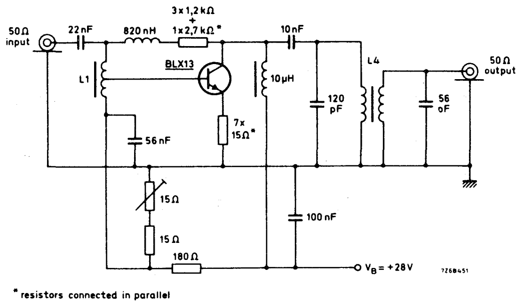


Fig.11.



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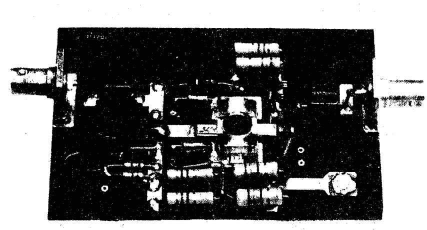


Fig.12. Single-stage driver module with BLX13.

Single-stage P.A. with 2 x BLY89A for 25-30 W P.E.P. and V_B = 12,5-13,5 V[13])

This amplifier (Fig.13) is intended for mobile or portable transmitters operating from a 12 V battery. Two BLY89A's are operated in Class-AB push-pull (the bias is obtained from a low internal resistance temperature-controlled circuit with two BD135's). A feature of this circuit is the inclusion of a computer-designed network for compensating the variation of the gain and input impedance of the transistors.

Other data for the amplifier:

i.m.d. up to max. output	⩽ −30 dB
overall 2-tone eff'y	⩾ 38 %
gain over the band	17,8 ± 0,6 dB
input VSWR	< 1,25.

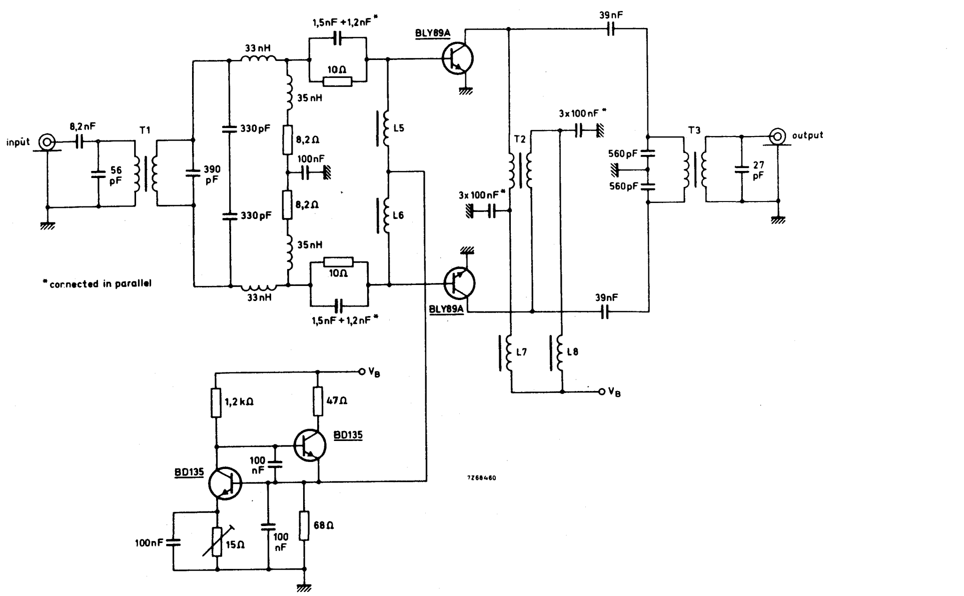


Fig.13.

Single-stage P.A. with 2 x BLX14 for 80-100 W P.E.P.[14])

This report contains a very detailed account of the design of a wide-band amplifier, including some remarks on linear operation. A theoretical amplifier is calculated and then a practical circuit (Fig.14) is given, with explanations where the practical circuit departs from the calculated design. This is an example of a computer-aided design.

Data for the amplifier:

supply voltage V_B	28 V
i.m.d. up to max. output	$\leqslant -30$ dB
overall 2-tone eff'y	$\geqslant 40$ %
gain over the band	16,8 ± 0,7 dB
input VSWR	< 1,4

The bias for the Class-AB operation is obtained from a circuit similar to that given in Reference 13.

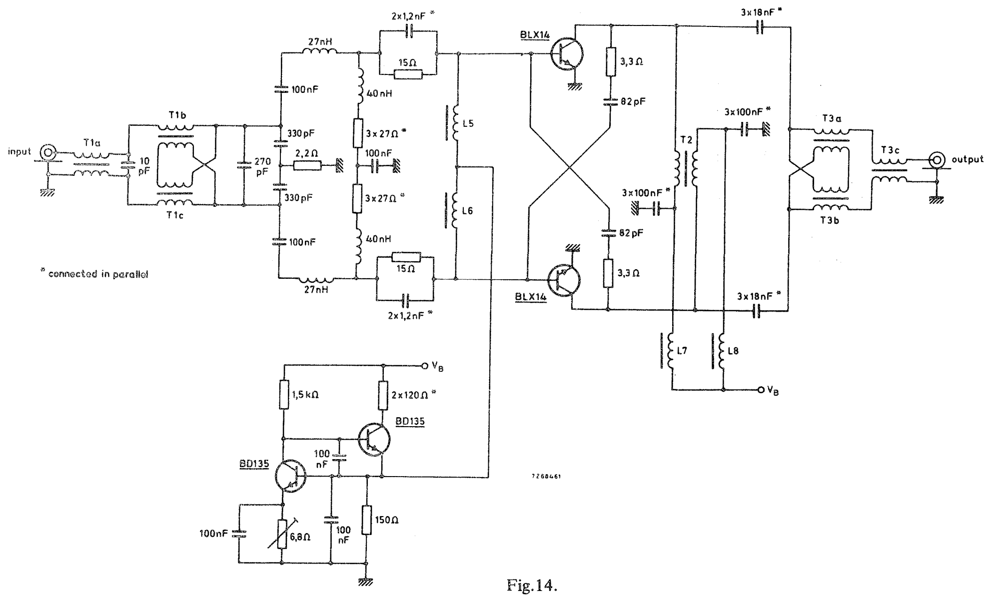
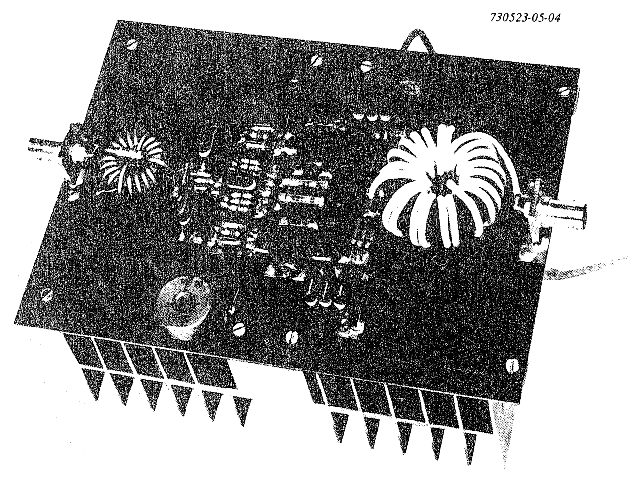


Fig.14.



Single-stage power amplifier with 2 x BLX14.

Single-stage P.A. with 2 x 545BLY for 165 W P.E.P. and V_B = 28 V[15])

This amplifier (Fig.15) follows the general design principles as given in Reference 14. The new transistor 545BLY (development number) comprises two BLX14 transistor elements in a single encapsulation.

Data for the amplifier:

supply voltage V_B	28 V
i.m.d. up to max. output	$\leqslant -30$ dB
overall 2-tone eff'y	> 33 %
gain over the band	12,9 ± 0,6 dB
input VSWR	$< 1{,}33$

The Class-AB bias is obtained from a temperature-compensated unit using a BD433 and a BD233.

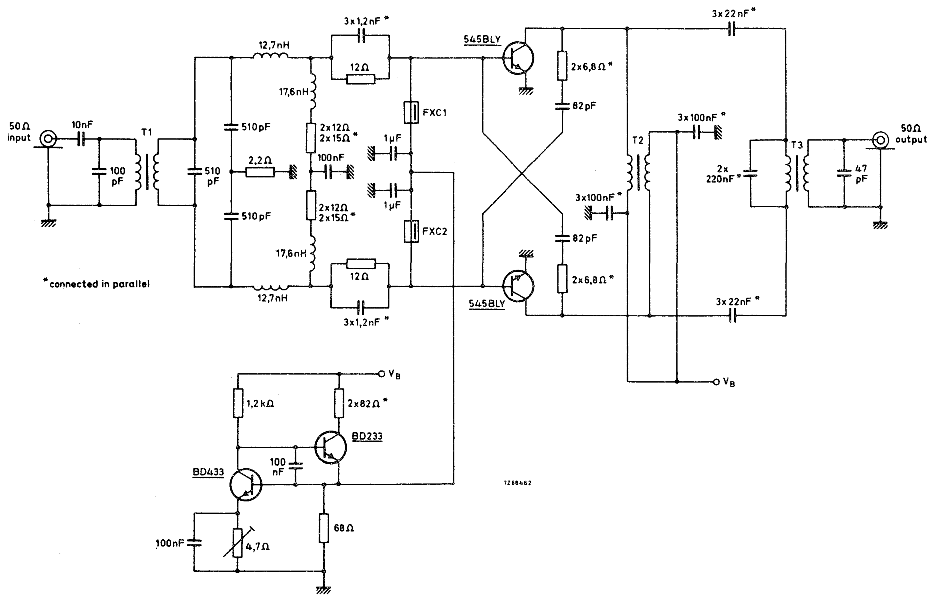


Fig.15.

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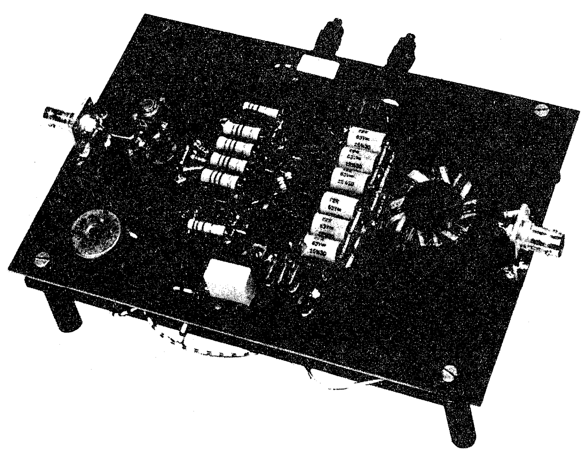

Single-stage power amplifier with 2 x 545BLY.

Single-stage P.A. with 2 x BLX15 for 300 W P.E.P. and V_B = 50 V [16])

This report gives results of an amplifier (Fig.16) using two BLX15 in Class-AB push-pull. As with the previous amplifier, this circuit is also designed in accordance with the procedure given in Reference 14. For this reason, the report deals chiefly with the differences between this and the earlier circuits.

Data for the amplifier:

i.m.d. up to max. output	⩽ −30 dB
gain over the band	16,8 ± 0,5 dB
input VSWR	< 1,2

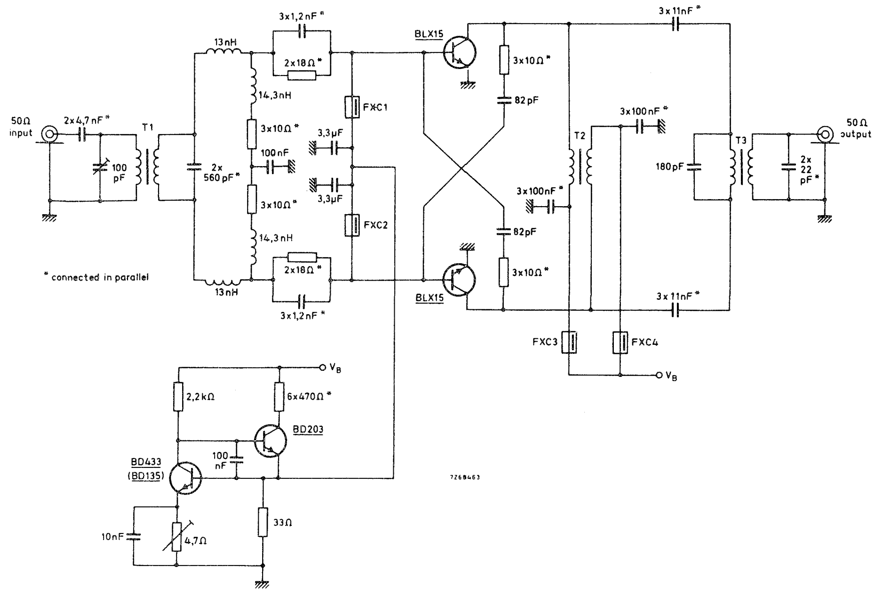

Fig.16.

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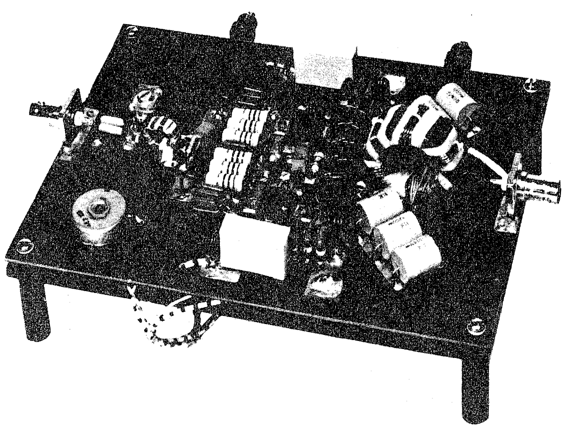


Single-stage power amplifier with 2 x BLX15.

Single-stage Driver for the S.S.B. Tube YL1230 using 2 x BLX13 in Class-A[16])

This amplifier was developed for driving the s.s.b. tube YL1230 up to 1 kW P.E.P. over the frequency band 1,6-28 MHz. The push-pull BLX13's operate in Class-A from a 28 V supply. The circuit diagram (Fig.17) shows two alternative outputs, one with, and one without a low-pass network. When the amplifier is used as a driver stage for the YL1230, the low-pass filter is needed to adapt the input capacitance of the tube. Alternatively, the amplifier can be used without the filter (i.e. with 50 Ω output) as a driver for a 300 W P.E.P. transmitter. When used with the YL1230 the required drive power for the BLX13's is 142 mW P.E.P. ± 0,7 dB and the input VSWR is below 1,6. The input voltage for the tube has an i.m.d. ⩽ −45 dB. When used with the 50 Ω output, the amplifier can deliver 14 W P.E.P. at an i.m.d. ⩽ −41 dB. In this case, the gain is 18,7 ± 0,15 dB and the input VSWR is below 1,5.

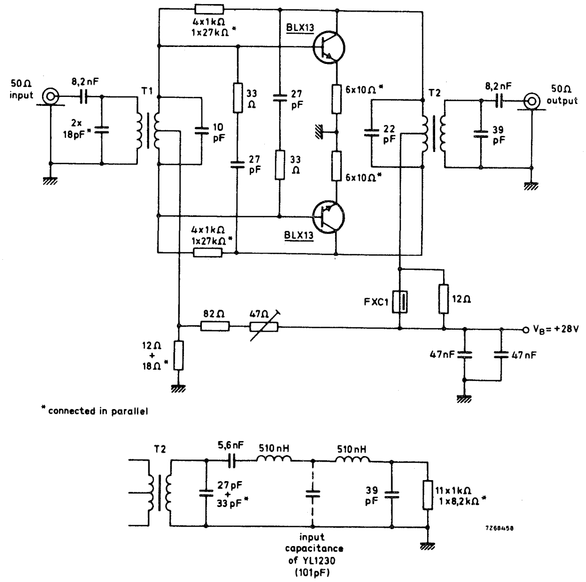
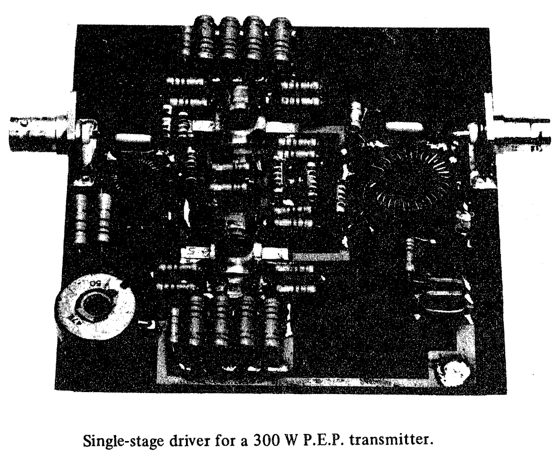


Fig.17.

730730-02-07



Single-stage driver for a 300 W P.E.P. transmitter.

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[2]) A.A. Bos, R.F. Zwanen, The BLY93A as a linear Class-A TV amplifier for Band III. Internal Report ECO7303.

[3]) A.A. Bos, Frequency tripler (157-470 MHz) with the BFR63. Internal Report ECO7214.

[4]) M.J. Köppen, Electronic antenna switches for the 160 MHz Band using BA182 diodes. Internal Report ECO6918.

[5]) M.J. Köppen, A three-stage 15 W power amplifier for the 470 MHz Band operating from a 12,5 V supply. Internal Report ECO7020.

[6]) M.J. Köppen, A single-stage power amplifier for 40 W output in the 470 MHz Band using 2 pieces BLX69 in parallel (V_B = 13,5 V). Internal Report ECO7216.

[7]) M.J. Köppen, A two-stage power amplifier for at least 100 W output at 175 MHz with 2 pieces BLY90 in the final operating from 12,5 V and 13,5 V supply voltage. Internal Report ECO7106.

[8]) M.J. Köppen, A two-stage wide-band power amplifier for 12 W minimum output in the 160 MHz Band, including a description of the swept-frequency equipment needed for alignment. Internal Report ECO6916.

[9]) H. v. Nugteren, An investigation on the behaviour of the BLY91A at lower power levels and frequencies. Internal Report ECO7205.

[10]) A. Walraad, A 10 MHz 25 W P.E.P. linear power amplifier using BDY92 at 20 V power supply. Internal Report ECO7110.

[11]) A. Walraad, A 10 MHz power amplifier using BDY92. Internal Report ECO7105.

[12]) M.J. Köppen, Single-stage wide-band (1,6-28 MHz) S.S.B. driver modules with BLY92A and BLX13 operating in Class-A. Internal Report ECO7113.

[13]) H. v. Nugteren, A single-stage wide-band (1,6-28 MHz) linear power amplifier for 25-30 W P.E.P. equipped with two pieces BLY89A and operating from a supply voltage of 12,5-13,5 V. Internal Report ECO7203.

[14]) J. Mulder, A single-stage wide-band linear power amplifier for 80-100 W P.E.P. in the 1,6-28 MHz frequency band equipped with two pieces BLX14. Internal Report ECO7114.

[15]) H.v. Nugteren, A single-stage wide-band (1,6-28 MHz) linear power amplifier for 165 W P.E.P. equipped with two pieces 545BLY and operating from a supply voltage of 28 V. Internal Report ECO 7305.

[16]) M.J. Köppen, A single-stage wide-band (1,6-28 MHz) linear power amplifier for 300 W P.E.P. using 2 x BLX15. Internal Report ECO7308.

[17]) M.J. Köppen, A wide-band (1,6-28 MHz) driver stage for the S.S.B. tube YL1230 equipped with two BLX13's in Class-A. Internal Report ECO7201.

Quick Reference Data for Semiconductors Mentioned in the Survey

BLX13

QUICK REFERENCE DATA

operation	class	V_{CE} (V)	f_1 (MHz)	f_2 (MHz)	P_L (W)	G_p (dB)	d_3 (dB)	I_C (A)	η_{dt} (%)
s.s.b.	A	26	28,000	28,001	0-8 (PEP)	>18	< −40	typ. 1,2	–
s.s.b.	AB	28	28,000	28,001	25 (PEP)	>18	typ. −35	typ. 1,28	typ. 35

operation	class	V_{CC} (V)	f (MHz)	P_S (W)	P_L (W)	G_p (dB)	I_C (A)	η (%)	$\overline{z_i}$ (Ω)	$\overline{Y_L}$ (mA/V)
c.w.	B	28	70	typ. 0,5	25	typ. 17	typ. 1,49	typ. 60	0,53-j1,4	42,5-j54

BLX14

QUICK REFERENCE DATA

operation	class	V_{CC} (V)	f (MHz)	P_L (W)	G_p (dB)	d_3 (dB)	I_{CZS} (A)
s.s.b.	A	28	1,6 to 28	15 (PEP)	> 13	typ. −40	2,0
s.s.b.	AB	28	1,6 to 28	7,5-50 (PEP)	> 13	< −30	0,1
c.w.	B	28	70	50	> 7,5		
c.w.	B	28	30	50	typ. 16		

BLX15

QUICK REFERENCE DATA

operation	class	V_{CC} (V)	f (MHz)	P_L (W)	G_p (dB)	d_3 (dB)	I_{CZS} (A)
s.s.b.	AB	50	1,6 to 28	20 to 150 (PEP)	> 14	<−30	0,10
s.s.b.	A	40	1,6 to 28	typ. 30 (PEP)	> 14	<−40	2,5
c.w.	B	50	70	150	> 10	–	–
c.w.	B	50	108	150	typ. 7,5	–	–

BLX67

QUICK REFERENCE DATA

R.F. performance up to T_h = 25 °C in an unneutralized common-emitter class B circuit

mode of operation	V_{CC} (V)	f (MHz)	P_S (W)	P_L (W)	I_C (A)	G_p (dB)	η (%)	$\overline{z_i}$ (Ω)	$\overline{Y_L}$ (mA/V)
c.w.	13,8	470	typ. 0,15	1,5	typ. 0,17	typ. 10	typ. 65	–	–
c.w.	13,8	470	typ. 0,35	3,0	typ. 0,28	typ. 9,3	typ. 79	3,0+j5,0	27−j38
c.w.	12,5	470	< 0,35	2,5	< 0,31	> 8,5	> 65	–	–
c.w.	12,5	175	typ. 0,03	3,0	typ. 0,29	typ. 20	typ. 84	2,4−j3,8	35−j40

BLX68

QUICK REFERENCE DATA

R.F. performance up to T_h = 25 °C in an unneutralised common-emitter class B circuit.

mode of operation	V_{CC} (V)	f (MHz)	P_S (W)	P_L (W)	I_C (A)	G_p (dB)	η (%)	$\overline{z_i}$ (Ω)	$\overline{Y_L}$ (mA/V)
c.w.	13,8	470	< 2,0	7,0	< 0,78	> 5,4	> 65	–	–
c.w.	13,8	470	typ. 2,0	7,8	typ. 0,81	typ. 5,9	typ. 70	2,3+j6,3	50–j36
c.w.	12,5	470	< 2,2	7,0	< 0,86	> 5,0	> 65	–	–
c.w.	12,5	175	typ. 0,4	7,2	typ. 0,87	typ. 12,6	typ. 66	3,0+j0,5	90–j40

BLX69

QUICK REFERENCE DATA

R.F. performance up to T_{mb} = 25 °C in an unneutralised common-emitter class B circuit.

mode of operation	V_{CC} (V)	f (MHz)	P_S (W)	P_L (W)	I_C (A)	G_p (dB)	η (%)	$\overline{z_i}$ (Ω)	$\overline{Y_L}$ (mA/V)
c.w.	13,5	470	<8	20	<2,28	>4	>65	1,1+j4,9	190–j45
c.w.	12,5	470	<6,8	17	<2,09	>4	>65		

BLY87A

QUICK REFERENCE DATA

R.F. performance up to T_{mb} = 25 °C in an unneutralised common-emitter class B circuit.

mode of operation	V_{CC} (V)	f (MHz)	P_S (W)	P_L (W)	I_C (A)	G_p (dB)	η (%)	$\overline{z_i}$ (Ω)	$\overline{Y_L}$ (mA/V)
c.w.	13,5	175	< 1,0	8	< 0,85	> 9	> 70	2,75+j1,5	74–j18
c.w.	12,5	175	typ. 1,0	8	typ. 0,91	typ. 9	typ. 70		

BLY88A

QUICK REFERENCE DATA

R.F. performance up to T_{mb} = 25 °C in an unneutralised common-emitter class B circuit.

mode of operation	V_{CC} (V)	f (MHz)	P_S (W)	P_L (W)	I_C (A)	G_p (dB)	η (%)	$\overline{z_i}$ (Ω)	$\overline{Y_L}$ (mA/V)
c.w.	13,5	175	< 2,65	15	< 1,71	> 7,5	> 65	2,3+j2,5	120–j7,8
c.w.	12,5	175	typ. 2,65	15	typ. 1,86	typ. 7,5	typ. 65		

BLY89A

QUICK REFERENCE DATA

R.F. performance up to T_{mb} = 25 °C in an unneutralised common-emitter class A circuit.

mode of operation	V_{CC} (V)	f (MHz)	P_S (W)	P_L (W)	I_C (A)	G_p (dB)	η (%)	$\overline{z_i}$ (Ω)	$\overline{Y_L}$ (mA/V)
c.w.	13,5	175	<6,25	25	<2,64	>6	>70	1,7+j1,4	209+j13,7

BLY90

QUICK REFERENCE DATA

R.F. performance up to T_h = 25 °C in an unneutralised common-emitter class B circuit.

mode of operation	V_{CC} (V)	f (MHz)	P_S (W)	P_L (W)	I_C (A)	G_p (dB)	η (%)	$\overline{z_i}$ (Ω)	$\overline{Y_L}$ (mA/V)
c.w.	12,5	175	<15,8	50	<5,33	>5,0	>75	1,3+j1,6	270+j160

BLY91A

QUICK REFERENCE DATA

R.F. performance up to T_{mb} = 25 °C in an unneutralised common-emitter class B circuit.

mode of operation	V_{CC} (V)	f (MHz)	P_S (W)	P_L (W)	I_C (A)	G_p (dB)	η (%)	$\overline{z_i}$ (Ω)	$\overline{Y_L}$ (mA/V)
c.w.	28	175	<0,50	8	<0,44	>12	>65	1,8+j1,0	17-j20

BLY92

QUICK REFERENCE DATA

R.F. performance up to T_{mb} = 25 °C in an unneutralised common-emitter class B circuit.

mode of operation	V_{CC} (V)	f (MHz)	P_S (W)	P_L (W)	I_C (A)	G_p (dB)	η (%)	$\overline{z_i}$ (Ω)	$\overline{Y_L}$ (mA/V)
c.w.	28	175	< 1,5	15	< 0,83	> 10	> 65	1,4+j2,15	32-j28

BLY93A

QUICK REFERENCE DATA

R.F. performance up to T_{mb} = 25 °C in an unneutralised common-emitter class B circuit.

mode of operation	V_{CC} (V)	f (MHz)	P_S (W)	P_L (W)	I_C (A)	G_p (dB)	η (%)	$\overline{z_i}$ (Ω)	$\overline{Y_L}$ (mA/V)
c.w.	28	175	<3,1	25	<1,5	>9	>60	1,0+j1,2	57,7–j52,7

BFR63
BFR64

QUICK REFERENCE DATA

Collector-base voltage (open emitter; peak value)	V_{CBOM}		max.	40		V
Collector-emitter voltage (open base)	V_{CEO}		max.	25		V
Collector current (peak value)	I_{CM}		max.	500		mA
Total power dissipation up to T_{mb} = 60 °C; $f \geqslant 1$ MHz	P_{tot}		max.	3,5		W
Junction temperature	T_j		max.	150		°C
Transition frequency at f = 500 MHz			BFR63	BFR64		
I_C = 75 mA; V_{CE} = 20 V	f_T	>	1000	1200		MHz
Output power at f = 200 MHz						
I_C = 70 mA; V_{CE} = 20 V; d_{im} = −30 dB	P_o	typ.	150	150		mW
Power gain at f = 200 MHz						
I_C = 70 mA; V_{CE} = 20 V	G_p	typ.	16	16		dB

BDY90 to 92

QUICK REFERENCE DATA

			BDY90	BDY91	BDY92	
Collector-base voltage (open emitter)	V_{CBO}	max.	120	100	80	V
Collector-emitter voltage (open base)	V_{CEO}	max.	100	80	60	V
Collector current (peak value)	I_{CM}	max.	15	15	15	A
Total power dissipation up to T_{mb} = 75 °C	P_{tot}	max.	40	40	40	W
Collector-emitter saturation voltage						
I_C = 10 A; I_B = 1 A	$V_{CE\,sat}$	<	1,5	1,5	1,0	V
Fall time						
I_C = 5,0 A; I_B = $-I_{BM}$ = 0,5 A V_{CC} = 30 V	t_f	<	0,2	0,2	0,2	µs
Transition frequency at f = 5 MHz						
I_C = 0,5 A; V_{CE} = 5 V	f_T	typ.	70	70	70	MHz

545BLY

QUICK REFERENCE DATA

mode of operation	class	V_{CC} (V)	f (MHz)	P_L (W)	G_p (dB)	d_3 (dB)	I_{CZS} (A)
s.s.b.	AB	28	1,6-28	10-85 (PEP)	>13	−30	0,2
c.w.	B	28	70	85	8		

SURVEY OF EARLIER ISSUES

Since 1964 the following publications have been issued in this series:

*510	July 1964	Two selective calling systems for mobile radio communication service
*511	June 1964	Mobile transceivers for the 80 Mc/s and 160 Mc/s bands
*512	July 1964	Portable transceivers for the 80 Mc/s and 160 Mc/s bands
*513	October 1964	Mobile and portable transceivers for the 470 Mc/s band
*514	September 1964	Mobile and portable transceivers for the 27 Mc/s band (Citizens Band)
*515	June 1964	Frequency multipliers with varactor diodes – theory and applications of the BAY66
*516	July 1964	A transistor i.f. amplifier for microwave link equipment
*517	October 1964	Investigations on a transistor aerial amplifier for TV Band III
*518	September 1965	A 600 mW, 165 Mc/s portable transmitter equipped with silicon planar transistors
*519	September 1965	Frequency doubler circuit, 500 Mc/s-1000 Mc/s
*520	June 1966	Mobile transmitters with the quick-heating tube YL1190
*521	June 1966	Flash-heating in mobile transmitters for the 160 and 470 MHz bands
*522	September 1966	Radar system using spin-tuned magnetrons
523	May 1967	A 165 MHz mobile transceiver (receiver part) using silicon transistors
524	March 1968	R.F. power amplifiers
525	August 1968	Transistor equipped aerial amplifiers
526	July 1968	All solid state mobile transceivers for the 160 MHz band using silicon-planar epitaxial overlay transistors
527	November 1968	TAB101 integrated four-transistor modulator
528	January 1969	A camera tube amplifier with FET input
529	January 1969	Principle and applications of junction circulators
530	June 1970	Design of h.f. wideband power transformers
531	December 1970	Solid state power amplifiers for the 160 MHz band using emitter-grid transistors
532	June 1971	Data transmission system using 30-series high noise immunity logic modules
533	November 1971	A 50 W d.c./d.c. converter with stabilized output voltage

* Out of stock; dye-line prints available on request.

Electronic components and materials

PHILIPS

laboratory report

central application laboratory CAB
eindhoven - the netherlands

number: ECO 7701 date: 2.3.1977

project: 6613 pages: A1, S1, R19

title DESIGN OF A SEMI-WIDE BAND POWER AMPLIFIER (146 - 174 MHZ) WITH BFQ43 AND BLW31

author M.J. Köppen

ABSTRACT

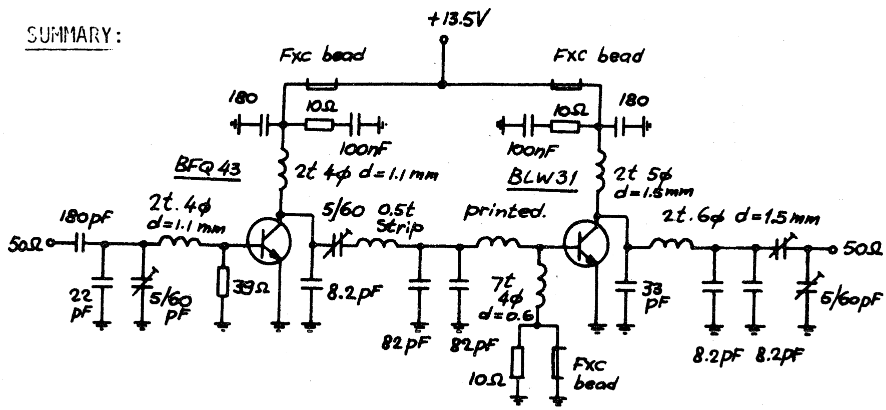
The VHF power transistors BFQ43 and BLW31 have been designed for the application in VHF communications equipment operating from an average voltage of 13.5V. The driver BFQ43 in T039 envelope has the emitter internally connected to the case, whilst the final transistor BLW31 has a ceramic envelope with 3/8 inch stud.

Five equal semi-wide band modules have been built with the BFQ43, BLW31 combination. They are tuneable between 146 and 174 MHz and deliver at least 28W output power for less than 250mW input. The amplifiers are stable for an output VSWR of 5 (0-360^{0}) for the range from zero to nominal output power of 28W by varying the drive power at a supply voltage of 13.5V.

A.H. Hilbers

Advies Octrooi d.d. 14 mrt. 1977	AV	GV		B......		BL
Opgave Mamo d.d. 22 mrt. 1977	AV	GV	SP	B		BL
Datum: -3 mrt. 1977	Mamo					

SUMMARY:



The figure shows the basic circuit of the prototype with both stages operating in class B. For the calculation of the matching networks the complex input- and load impedances of the transistors have been calculated with the aid of a computer programme (C.A.B. reports ECO 7112 and 7118).

The average values are:

BFQ43: R_i = 3,12Ω; X_i = -1,80Ω; R_L = 21,34Ω; C_L = -25,8pF

BLW31: R_i = 0,94Ω; X_i = +1,31Ω; R_L = 2,84Ω; C_L = 20,6pF

To reduce the costs and to simplify the lining-up the number of tuning elements has been reduced as much as possible. Furthermore the circuit contains a number of components to guarantee stability under mismatch conditions.

The amplifier is printed on a double sided copper clad epoxy glass-fibre board that is housed in a tinned box. A small intermediate heatsink is applied that has to be screwed to an external one.

Some typical results are: (P_o = 28W; V_B = 13,5V).

f (MHz)	P_i(mW)	P_r(mW)	I_t(A)	eff(%)	gain (dB)
146	225	0	3.60	58	20.95
153	200	0	4.00	52	21.46
160	190	0	3.95	53	21.68
167	185	0	4.05	51	21.80
174	195	20	4.30	48	21.57

Central application laboratory CAB Eindhoven - The Netherlands

1. Introduction

The BFQ43 (dev. nr. 672BLY) and BLW31 (dev. nr. 673BLY) are VHF power transistors designed for the application in fixed or mobile communications equipment where they operate from an average voltage of 13.5V.

The devices have high power gain what contributes to simplification of the construction of the r.f. power part of this equipment.

The BLW31 is encapsulated in a ceramic envelope with 3/8 inch stud (NO 181/A2) whilst the BFQ43 has its crystal mounted in a SOT 5/11 (TO39) envelope on a beryllia chip with the emitter internally connected to the case. So, it is possible to solder the case to the printed wiring board.

This report gives an application of both devices in a class B amplifier, operating from a supply voltage of 13.5V. The amplifier is tuneable and covers all channels between 146 and 174 MHz. The output power amounts to 28W for less than 250mW input power.

For an output VSWR of 5 ($0-360^{0}$) the amplifier is stable for the range from zero to nominal output power of 28W by varying the drive power. The supply voltage is 13.5V.

2. Circuit description

Fig. 1 shows the basic circuit of the prototype.
It is a two stage amplifier with both stages operating in class B.

Electronic components and materials

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To reduce the costs of the amplifier and to simplify the lining up it has been attempted to reduce the number of tuning elements as much as possible.

2.1. The output network of the BLW31

Starting originally from the more complex Chebishev low-pass impedance transforming network between the collector of the BLW31 to the 50 Ω load, with only one tuning element, it appeared that it was rather difficult to obtain a stable amplifier for load mismatch.

So, ultimately, we have chosen for the conventional method with two tuning elements.

For calculation of these networks the value of the R_L and C_L has to be known. They have been calculated for Po = 28W and V_b = 13.5V at a number of spot frequencies.
Moreover the complex input impedance and gain are shown (for L_e = 0,4 nH) in the following list.

	parallel		series		
f (MHz)	R_L(Ω)	C_L(pF)	R_i(Ω)	X_i(Ω)	gain (dB)
146	2.76	11.9	0.95	1.08	11.70
153	2.81	15.9	0.94	1.20	11.34
160	2.84	20.8	0.94	1.31	10.98
167	2.88	25.9	0.94	1.43	10.64
174	2.92	28.5	0.93	1.54	10.33
av.	2.84	20.6	0.94	1.31	Δ gain = 1.37 dB

A valuable aid in calculating these values was the computer program as described in C.A.B. reports ECO 7112 (ref. 1) and ECO 7118 (ref. 2) or in Electronic Applications (refs. 3 and 4).

From the table it may be concluded that for the high power device BLW31 the impedance transformation factor to 50Ω is rather high.

A method to solve the problem is to return a part of the r.f. current to ground by fixed capacitors ($C_{13,\ 14}$) before variable matching is applied. It corresponds to an impedance transformation in two steps and improves the overall gain and efficiency.

Fig. 2a, b show the applied circuit configuration.

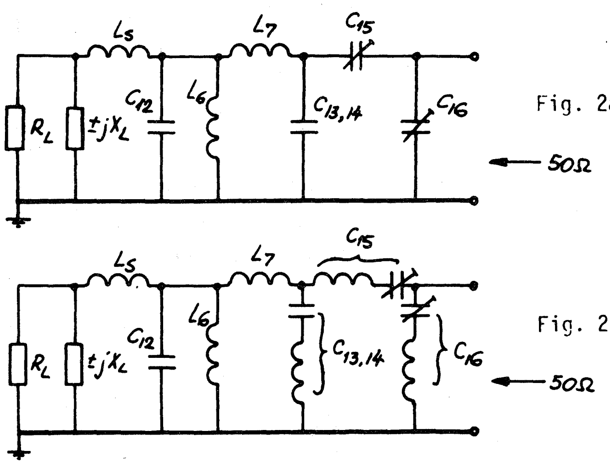

Fig. 2a.

Fig. 2b.

L_6 serves as an r.f. choke for the supply of the direct voltage to the collector. The impedance of this choke is about 5 to 7 times that of the collector load impedance.

Together with the r.f. bypass capacitor C_{11} (180pF) and the series connection of R_4 (typically 10Ω) and C_{10} (typically 100nF) it forms a broadband collector impedance network (Fig. 1) that contributes to the stability when the external load is mismatched.

The supply choke F_3 gives a better definition of the minimum impedance of the supply source, what is specially important to suppress instability when the units are operated from certain types of electronically stabilized power supplies.

Because the often applied ferroxcube chokes of the VK200 type (4312 020 36640) are somewhat bulky in this compact amplifier, smaller ones (4312 020 15171) are used. They have been realized with 3 turns of 0.6 mm CuEm. wire.

In Fig. 2 the collector series inductance (internally and externally) is incorporated in L_s.

The capacitor C_{12} was added during the experiments. It decreases the collector peak voltage to a value below the point where avalanching may occur and so improves the stability. The value is 33pF.

As Fig. 2b shows, the capacitors have series inductance which may not be neglected in circuit design for the VHF range. In case of tuning capacitors (here 60 pF) we have calculated with 5nH and for fixed capacitors with 8nH per component.

2.2. The interstage network BFQ43 → BLW31

To get an idea of the requirements of the set-up of the interstage network the calculated performance of the driver-stage is shown in the following list:

$P_o = 4W$, $V_b = 13.5V$, $R_{be} = 39\Omega$

	parallel		series		
f(MHz)	$R_L(\Omega)$	C_L(pF)	$R_i(\Omega)$	$X_i(\Omega)$	gain (dB)
146	22.12	-28.5	3.17	-2.50	14.95
153	22.01	-27.0	3.13	-2.13	14.60
160	21.76	-25.6	3.12	-1.78	14.25
167	21.73	-24.5	3.09	-1.46	13.93
174	21.59	-23.3	3.08	-1.14	13.60
av.	21.84	-25.8	3.12	-1.80	Δ gain = 1.35 dB

The interstage network has to match the parallel connection of average 21.84Ω and 25.8pF to the average input impedance (0,94 + j 1,31Ω) of the BLW31.

It had to be realized with a minimum number of tuning elements.
After some experiments the configuration of Fig. 3 was applied.

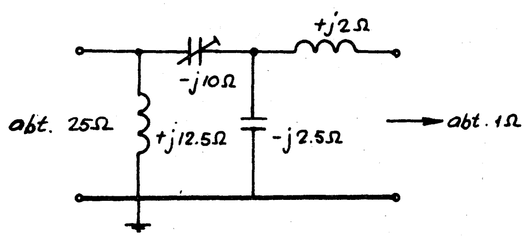

Fig. 3.

The circuit is an equivalent for the solution of a corresponding problem in the VHF modules (BGY32, 33, 35 and 36). Fig. 3 shows the practical circuit.

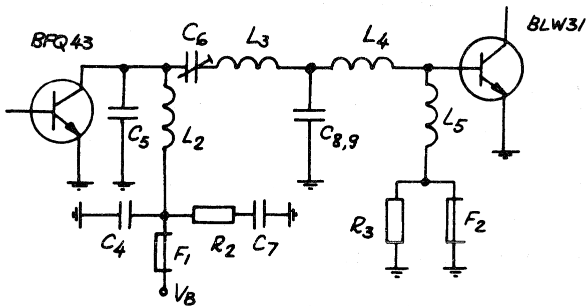

Fig. 4.

The function of the elements is:

L_2 is the r.f. choke for the direct voltage supply to the collector of the driver stage.
According to Fig. 3 the value of L_2 plays a role in the configuration, so the rule that the value has to be 5-7 times the collector load impedance is not valid in this case.

C_4 and R_2 in series with C_7 have equal values as C_{11}, R_4 and C_{10} because their function is the same.

The reactance of C_6 is average -j10 Ω , what in practice means a trimmer of 60pF.
From the components list it can be seen that L_3 is a U shaped inductance, whilst C_8 and C_9 in parallel are connected to the tap on the inductance L_3, L_4.

It must be noted that the leads of C_8 and C_9 (82pF) have to be very short. The tapping point is critical, what means that it is possible to influence the frequency response and matching strongly when this point is shifted.
To prevent positive feedback the earth current of $C_{8,9}$ and the filter capacitor $C_{13,14}$ have been led to different emitter leads of the final transistor.

The impedance of the combination L_5, R_3 and F_2 does not really influence the circuit. It consists of a carefully chosen combination that prevents parasitic oscillations under severe mismatch conditions. The ferrite bead choke F_2 across R_3 is added because it is not advisable that the base current causes the transistor to operate in class C. This network gives a better suppression of parasitics than a single ferroxcube choke.

The capacitor C_5 has the same function as C_{12} in the final stage.

2.3. The input circuit of BFQ43

From the table mentioning the impedances one can see the complex input impedance of the BFQ43 over the range of 146-174 MHz, which has to be matched to 50 Ω by the input circuit. The ohmic part is rather constant, but the imaginary part X_i varies more than a factor 2. This means that at least one variable element in the input section is needed.

The calculation of the component values goes in a conventional way when one starts from the mentioned average values.

Fig. 5 shows the actual L type circuit.

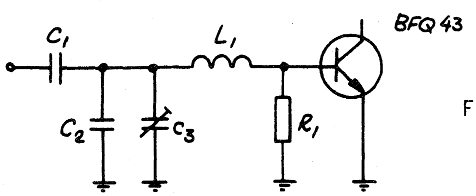

Fig. 5.

A base return resistor $R_1 = 39\,\Omega$ forms a good compromise between constancy of power gain and input impedance between 146 and 174 MHz. Application of a single resistor contributes to the stability of this stage when the final or (and) the input are mismatched.

C_3 is the variable capacitor of 60pF. The blocking capacitor C_1 is 180pF. This value has been chosen because it shows series resonance around 160 MHz.
If one wants to apply another style of capacitor it is advisable to chose the value (the leads short-circuited) with the aid of, for example, a grid-dip meter.

Of course, the same yields for the decoupling capacitors C_4 and C_{11}.

3. Constructional details

A suitable printed-wiring board design is given in Fig. 6 whilst the components lay-out is drawn in Fig. 7a.

It has been attemped to miniaturize this design somewhat to make it suitable for practical mobile equipment.

The board is of 1.5 mm (1/16 in) double sided copper-clad epoxy glass-fibre with copper thickness of appr. 35 μ.
Soldered rivets (2 mm) are inserted to make ground connections between the upper and lower sides of the board.

As Fig. 7b shows a small intermediate aluminium heatsink has been applied. This one is too small to handle the power dissipation. So it has to be screwed to a suitable one that can dissipate at least power levels between 20 and 25W under non-mismatched conditions.

In the laboratory set-up it was screwed to a water-cooling unit.

The p.c. board is housed in a tinned box being soldered to the lower copper sheet.
It is permitted to solder the case of the BFQ43 directly to the lower sheet, whilst the heatsink is pressed against the cap. For a good heat-transfer some silicone heatsink compound (Dow Corning 340) is used.

For the stud type BLW31 a square hole has been choosen instead of a round one. The advantage is that the direct contact from upper to lower sheet under the emitter leads can be made with wide copper strip instead of hollow rivets.

Fig. 8 shows the construction drawing of the intermediate heatsink.

4. Mismatch tests and spurious generation

It will be clear that one of the most important requirements is that the amplifier remains stable and survives without damage or degradation of the active devices when mismatch at the output is applied.

The stability tests have been done according to the following specifications:

- measuring frequencies: 146, 153, 160, 167 and 174 MHz
- supply voltage V_b = 13.5V
- output power Po = 28W
- drive level 0 - 250 mW over 50Ω
- VSWR (output) 1:5 (0-360^0)
- heatsink temperature appr. 20^{0}C.

Realization of stability under mentioned conditions was not so easy.
The greater part of the development time was spent on this subject and several tests have been done.
So, it is advisable to follow the circuit diagram and well-considered components lay-out as close as possible.

Tests have been made with the set-up of Fig. 9, in which the spectrum analyzer HP 8558 B appeared to be an indispensable instrument. To be sure of 50 Ω source impedance a 3dB attenuator pad was inserted in the chain of the signal generator type SMLU (Rhode and Schwarz).

Input power and VSWR where measured with the NAU.

Fig. 10 shows the composition of the VSWR = 5 unit and Fig. 11 the diagram of the reactance unit.

Resuming, it can be said that a reasonable stability has been reached under normally tuned conditions.

5. Measurements

For demonstration purposes totally 5 amplifiers have been constructed. They are numbered 1, 2, 3, 5 and 6.
The interstage and output network are tuned for maximum output power and the input section for minimum reflection.

The results are:

nr. 1 Po = 28W; V_B = 13.5V

f(MHz)	P_i(mW)	P_r(mW)	I_t(A)	eff(%)	gain (dB)
146	245	0	3.90	53	20.58
153	225	0	3.90	53	20.95
160	210	0	3.90	53	21.25
167	190	0	4.35	48	21.68
174	195	30	4.15	50	21.57

nr. 2 Po = 28W; V_B = 13.5V

f(MHz)	P_i(mW)	P_r(mW)	I_t(A)	eff(%)	gain (dB)
146	225	0	3.60	58	20.95
153	200	0	4.00	52	21.46
160	190	0	3.95	53	21.68
167	185	0	4.05	51	21.80
174	195	20	4.30	48	21.57

nr. 3 Po = 28W; V_B = 13.5V

f(MHz)	P_i(mW)	P_r(mW)	I_t(A)	eff(%)	gain (dB)
146	210	0	3.95	53	21.25
153	195	0	3.90	53	21.57
160	175	0	3.90	53	22.04
167	150	0	3.70	56	22.71
174	170	0	4.20	49	22.18

nr. 5 Po = 28W; V_B = 13.5V

f(MHz)	P_i(mW)	P_r(mW)	I_t(A)	eff(%)	gain (dB)
146	230	0	4.10	51	20.85
153	200	0	4.10	51	21.46
160	170	0	3.95	53	22.18
167	145	0	4.00	52	22.86
174	200	10	4.25	49	21.46

nr. 6 Po = 28W; V_B = 13.5V

f(MHz)	P_i(mW)	P_r(mW)	I_t(A)	eff (%)	gain (dB)
146	205	0	4.10	51	21.35
153	195	0	3.65	53	21.57
160	175	5	3.70	57	22.04
167	170	0	3.85	54	22.17
174	170	10	3.80	55	22.17

All units are stable for load mismatches up to a VSWR of 5 (any phase) for each input drive up to an output power of 28W.

The collector current of the BFQ43 amounts to 300 - 400 mA.

6. References

Ref. 1: A.H. Hilbers - Calculation of the large-signal behaviour of R.F. power transistors by means of an equivalent circuit. Part I. C.A.B. report ECO 7112.

Ref. 2: J. Mulder - Calculation of the large-signal behaviour of R.F. power transistors by means of an equivalent circuit. Part II. C.A.B. report ECO 7118.

Ref. 3: A.H. Hilbers - Large-Signal behaviour of r.f. power transistors. Part 1, Analysis of the equivalent circuit, Electronic applications bulletin, volume 31, number 3, 1972, p.p. 135 - 150.

Ref. 4: J. Mulder - Large-signal behaviour of r.f. power transistors.
Part 2, Computer programme.
Electronic applications bulletin,
volume 31, number 4, 1972,
p.p. 218 - 233.

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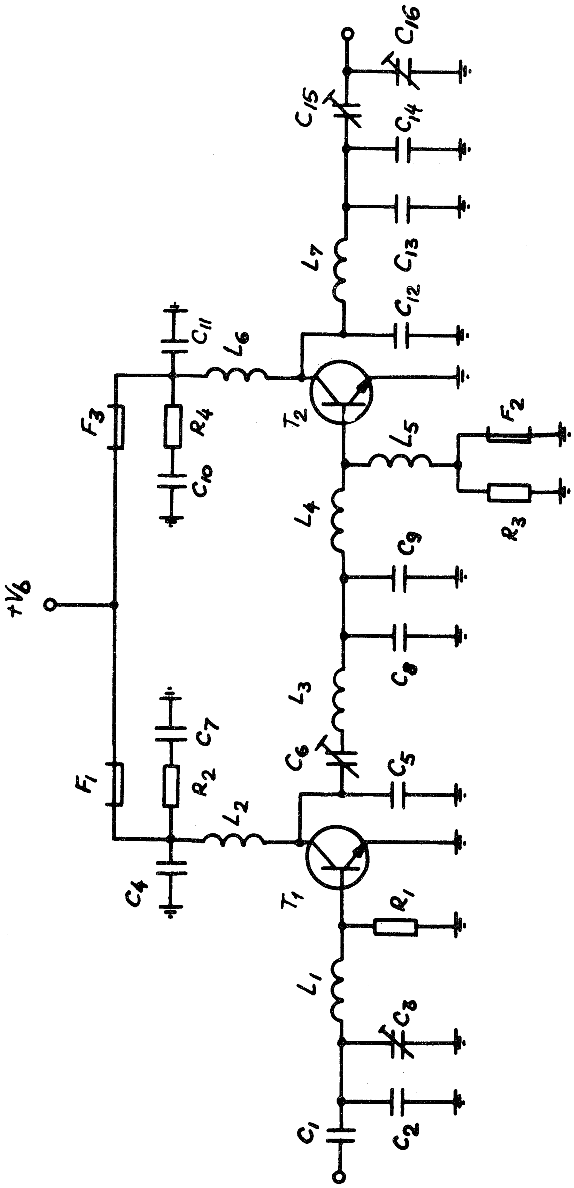

Fig. 1.

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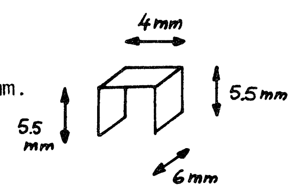
7. Parts list (Figs. 1, 7a)

C_1, C_4, C_{11} = 180pF, 500V d.c.	2222 655 03181
C_2 = 22pF ± 2%, 500V d.c.	2222 650 10229
C_3, C_6, C_{15}, C_{16} = 5/60 pF, film dielectric trimmer	2222 809 08003
C_5, C_{13}, C_{14} = 8.2pF ± 0.25pF, 500V d.c.	2222 650 09828
C_7, C_{10} = 100nF polyester ± 10%	2222 342 45104
C_8, C_9 = 82pF ± 2%, 500V d.c.	2222 650 58829
C_{12} = 33pF ± 2%, 500V d.c.	2222 650 10339
R_1 = 39 Ω, carbon ± 5%, CR25 style	2322 101 33399
R_2, R_3, R_4 = 10 Ω, carbon ± 5%, CR25 style	2322 101 33109
F_1, F_2, F_3 = ferroxcube bead with 3 turns of 0.5 or 0.6 mm CuEm. wire.	4313 020 15171

L_1 = 2 turns, $D_{int.}$ = 4.0mm, d = 1,1 mm CuEm. wire, spacing = wire diameter, leads 2 x 4 mm.

L_2 = 2 turns, $D_{int.}$ = 4.0mm, d = 1.1mm CuEm. wire, closely wound, leads 2 x 4 mm.

L_3 = U-shaped copper strap, width 6 mm. thickness 0.1 mm.



L_4 = printed (see p.c. board)

L_5 = 7 turns, $D_{int.}$ = 4.0 mm, d = 0.6 mm CuEm. wire, closely wound, leads 2 x 4 mm.

L_6 = 2 turns, $D_{int.}$ = 5.0 mm, d = 1.5 mm CuEm. wire, closely wound, leads 2 x 4 mm.

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L_7 = 2 turns, $D_{int.}$ = 6.0 mm, d = 1.5 mm CuEm wire, spacing appr. 0.5 mm, leads 2 x 5 mm.

T_1 = BFQ43 9334 120 90XXX
T_2 = BLW31 9334 120 20XXX

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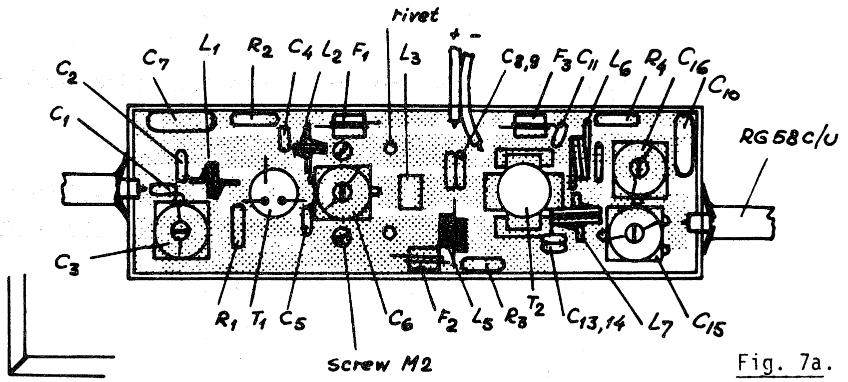


Fig. 7a.

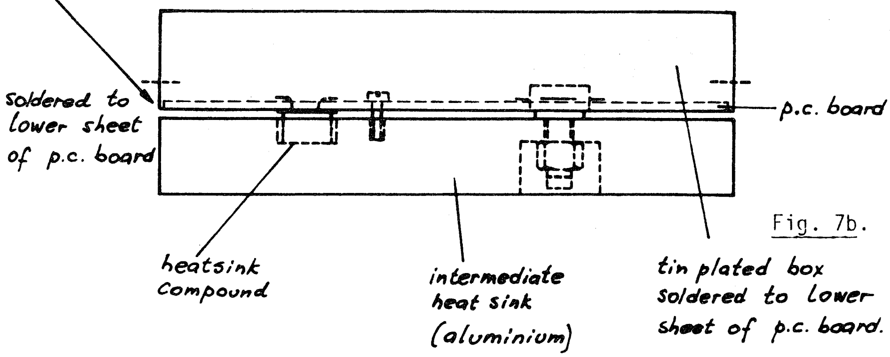
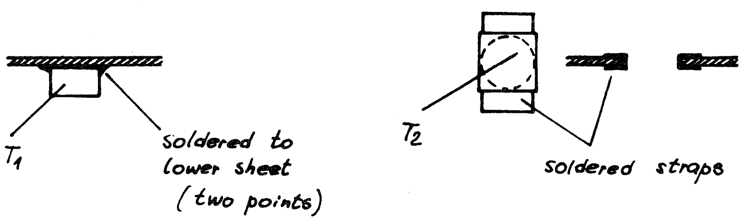


Fig. 7b.



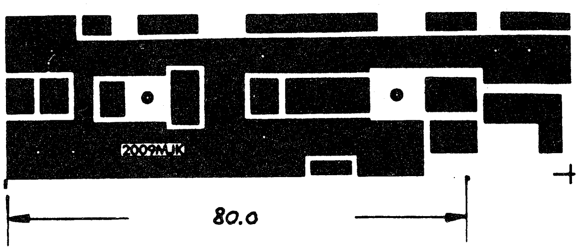


Fig. 6.

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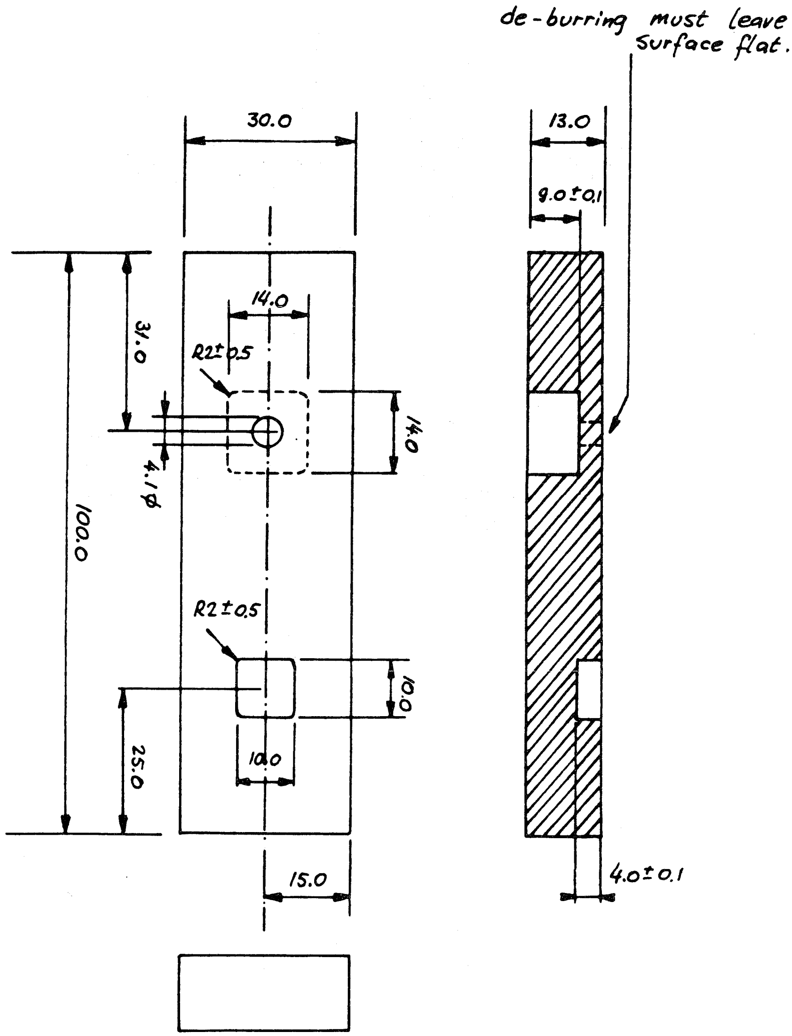

Tolerances not specified : ± 0.2 mm

Material : aluminium

HEAT SINK

Fig. 8.

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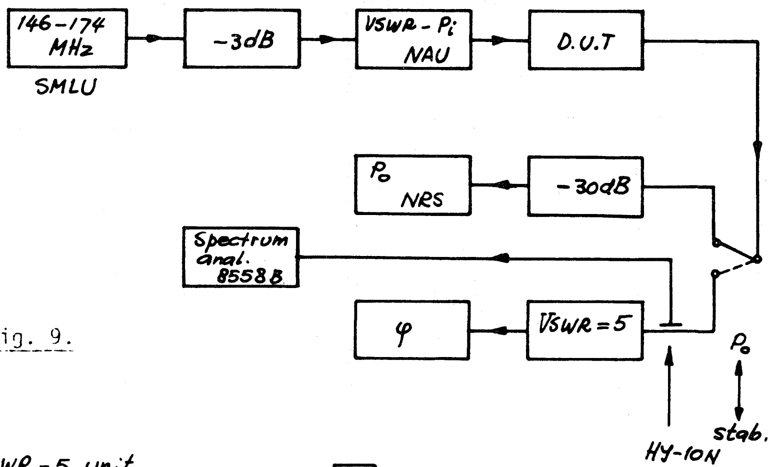

Fig. 9.

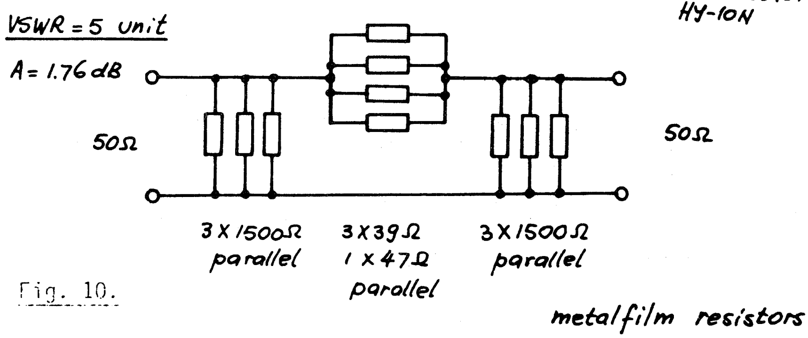

Fig. 10.

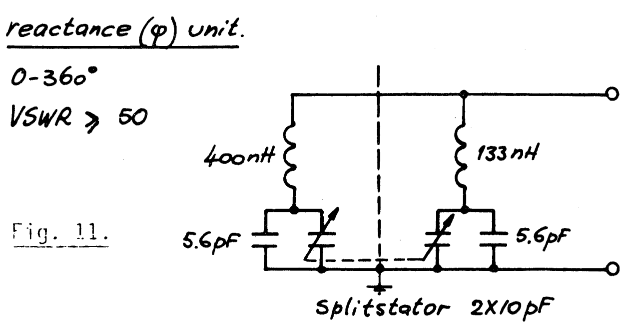

Fig. 11.

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laboratory report

central application laboratory CAB
eindhoven - the netherlands

number:	ECO 7703	date:	16.05.1977
project:	6706	pages:	A1 + S1 + R8

title POWER TRANSFORMERS FOR THE FREQUENCY RANGE OF 30-80MHZ

author A.H. Hilbers

ABSTRACT

In this report design information is given for transformers with a power handling capability up to 300W in the frequency range of 30-80MHz.

The most suitable core material is ferrite type 4C6.
The efficiency of these transformers is typically 98%.

A.H. Hilbers
appr. R.A. Pölzl

Advies Octrooi d.d. 22 jun. 1977	X AV	GV		B......		BL	
Opgave Mamo d.d. 22 jun. 1977	X AV	X GV	X SP	B..		BL	
Datum: 15 jun. 1977	Mamo						

SUMMARY

In this frequency range only transmission line transformers can be used. For the windings coaxial cables with P.T.F.E. isolation are recommended.

The size of the core is based on a 1% power loss and a dissipation of $350mW/cm^3$ corresponding with a flux density of 6 Gauss at 80MHz.

The required number of turns is determined by the ratio $R_p/L = 860\Omega/\mu H$ in which R_p is the loss resistance and L the inductance in parallel with the input or output terminals.

In the appendix the relation between the above mentioned quantities is derived.

In the report a practical example is given of a symmetrical 1:4 impedance transformer with a power handling capability of 120W.

1. Introduction

In Ref. 1 information was given on the design of power transformers mainly intended for the frequency range of 1.6-28MHz. In this report some additional information will be presented for the frequency range of 30-80MHz.

2. Choice of core material

The best available ferrite for this frequency range is 4C6. In this material a series of toroids can be obtained in different sizes according to the table below.

D (mm)	d (mm)	h (mm)	A (mm^2)	A/l (mm)	V (mm^3)
36	23	15	97.7	1.06	8500
23	14	7	31.5	0.552	1790
14	9	5	12.5	0.351	445
9	6	3	4.51	0.193	105

Table I

In this table:
D = outside diameter
d = inside diameter
h = height
A = cross-section
l = average length of the lines of force
V = volume

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3. Power handling capability

An important question in the design of a power transformer is how much R.F. power can be handled by a given toroid. Restricting ourselves to the core losses at this moment it can be said that these losses are highest at the maximum frequency of operation i.e. 80MHz.

From practical experience we have found that a core dissipation of 350mW/cm^3 can be allowed without excessive rise of the core temperature. As it is a realistic target to keep the core losses below 1% of the power handled by the transformer we come to the following recommendations for the power handling of the different toroids (see table II).

D x d x h (mm^3)	P_{RF} (W)
36 x 23 x 15	300
23 x 14 x 7	60
14 x 9 x 5	15
9 x 6 x 3	3

Table II

The core dissipation of 350mW/cm^3 mentioned above corresponds with a flux density of 6 Gauss at 80MHz as can be found on page A43 of part 4a of the Philips Data Handbook: Components and Materials, dated October 1976.

4. Determination of the number of turns

In the frequency range of 30-80MHz the number of turns is entirely determined by the loss resistance in parallel with the input or output terminals of the transformer being caused by the core losses. According to the Appendix the core loss figures given in the previous section can be expressed in another way, viz:

$$\frac{R_p}{L} = \frac{\omega^2 B_{max}^2}{2\mu_o \mu_r} \cdot \frac{V}{P_L} ,$$

in which:

R_p = loss resistance in parallel with input or output terminals
L = inductance in parallel with input or output terminals
B_{max} = maximum flux density
μ_r = relative permeability being typ. 120 for 4C6 material
V = volume of transformer core
P_L = power loss in core

Using the figures given in section 3 we get:

$$\frac{R_p}{L} = 860\Omega/\mu H \quad \text{at } f = 80\text{MHz}.$$

This ratio is hardly dependent on the flux density and therefore it is very useful for defining the number of turns as will be shown by a practical example in section 6.

Applying the above mentioned criterion ensures a sufficiently high reactance in parallel with the input or output terminals at the lowest frequency of operation. So this reactance caused by the inductance of the winding needs no compensation.

5. Winding losses

In this frequency range conventional transformers can not be used because of their stray-inductance. The only suitable type is the transmission line transformer. For the windings we can choose a.o.:

- twisted enamelled copper wire
- miniature twin lead
- coaxial cable with P.T.F.E. isolation, available in several diameters and characteristic resistances.

The first and second are not recommended for high power operation. The third type is e.g. available in 50 ohms version with diameters of 1.7 and 2.8mm. Some important properties of these cables at 80MHz are given in the table below.

Outside diameter	1.7	2.8	mm
power loss	0.40	0.24	dB/m
power handling capability	100	200	W

Table III

At lower frequencies the power loss is less and the power handling capability higher.

6. Practical example

Suppose that in a 100W transmitter the output transistors are connected in push-pull. The optimum load impedance between the collectors is 12.5 ohms and this must be transformed to 50 ohms. Then we need a symmetrical 1:4 impedance transformer plus a balun. The first one will be worked out in detail. A schematic diagram is given in fig. 1.

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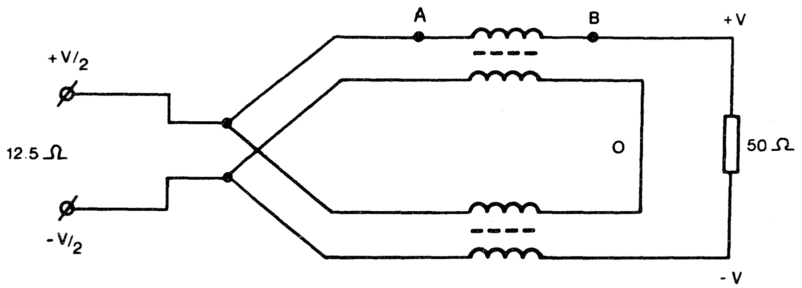

Fig. 1

From table II we see that this transformer can be made with one 36mm toroid but also with two 23mm toroids, in which case the power handling capability is still 120W. The latter solution is more attractive because it is smaller.

The optimum characteristic impedance for each winding is 25 ohms which can be realized by the parallel connection of two 50 ohms cables of 1.7mm diameter. As the power is transported through 4 cables, each cable is loaded with 25W being a quarter of the allowable maximum.

The number of turns will be calculated with the 50 ohms output terminals as a reference point.
To keep the core losses below 1% we must keep the parallel loss resistance above 5000 ohms.
This means an inductance of:

$$L = \frac{R_p}{860} = \frac{5000}{860} = 5.81\mu H \text{ (see section 4)}$$

Between the points A and B in fig. 1 the voltage is one quarter of the output voltage. This means that the inductance between these points must be one sixteenth of that across the output terminals, so:

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$$L_{AB} = \frac{L}{16} = \frac{5.81}{16} = 0.363\mu H.$$

Now the number of turns can be calculated with:

$$n = \sqrt{\frac{L1}{\mu_o \mu_r A}} = 1.48$$

In practice we will choose of course 2 turns by which the core losses reduce to:

$$\left(\frac{1.48}{2}\right)^2 . 1\% = 0.55\%$$

The inductance in parallel with the output terminals rises to:

$$\left(\frac{2}{1.48}\right)^2 \times 5.81 = 10.6\mu H$$

This corresponds to a reactance of 2000 ohms at 30MHz which is high enough to be neglected.

To realize the windings cables are required with a length of appr. 98mm giving a cable loss of 0.039dB or 0.91%.

So the total calculated loss of this transformer is:

$$0.55 + 0.91 = 1.46\% \text{ at } f = 80MHz.$$

Reference:
Application Information Nr. 530 "Design of H.F. Wideband Power Transformers" by A.H. Hilbers, June 17th, 1970.

APPENDIX

In the Data Handbook curves are giving showing the core losses expressed in KW/m^3 or mW/cm^3 versus the flux density B with the frequency as a parameter. It is often useful to know what this means in terms of an equivalent loss resistance in parallel with the inductance. The power dissipated in this resistance is equal to:

$$P_L = \frac{E_{max}^2}{2R_p} \qquad (1)$$

On the other hand:

$$B_{max} = \frac{E_{max}}{\omega A n} \qquad (2)$$

Eliminating E_{max} in (1) and (2) gives:

$$P_L = \frac{\omega^2 B_{max}^2 A^2 n^2}{2R_p} \qquad (3)$$

Further we know that:

$$L = \frac{\mu_o \mu_r n^2 A}{l} \qquad (4)$$

So that:

$$n^2 A = \frac{Ll}{\mu_o \mu_r} \qquad (5)$$

Substituting (5) in (3) we get:

$$P_L = \frac{\omega^2 B_{max}^2 A L l}{2\mu_o \mu_r R_p} \qquad (6)$$

The product Al is equal to the volume of the core V, so that:

$$\frac{P_L}{V} = \frac{\omega^2 B_{max}^2}{2\mu_o\mu_r} \cdot \frac{L}{R_p} \qquad (7)$$

or:

$$\frac{R_p}{L} = \frac{\omega^2 B_{max}^2}{2\mu_o\mu_r} \cdot \frac{V}{P_L}$$

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laboratory report

number : EC07801 date : 09-03-1978

title : A 2 STAGE 14W POWER AMPLIFIER WITH BFQ42/BLW29 TUNABLE FROM 146-174 MHz

author : A.BOEKHOUDT

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laboratory report

central application laboratory CAB
eindhoven - the netherlands

number: ECO 7801 date: 09.03.1978
project: 6701 pages: A1, S2, R13

title A 2-STAGE 14W POWER AMPLIFIER WITH BFQ42/BLW29 TUNABLE FROM 146-174 MHz

author A. Boekhoudt

ABSTRACT

The power amplifier described in this report is intended for application in transmitters of mobile radios. The number of amplifier stages could be reduced from 3 to 2 by applying the high-gain V.H.F. transistors BFQ42 and BLW29.

The output power is 14W at a supply voltage of 13.5V.
The tuning range is from 146 to 174 MHz using a number of tuning elements which is 50% less than in earlier designs.

The overall power gain and efficiency are 22.5 ± 1dB and 60% typ.

Parasitic oscillations do not occur up to an output VSWR of 4.4 (any phase) combined with any drive up to an output power of 14W.

R.A. Pölzl

Advies Octrooi d.d. 21 mrt. 1978	X AV	GV		B		BL
Opgave Mamo d.d. 21 mrt. 1978	X AV	X GV	X SP	B		BL
Datum: 10 mrt. 1978	Mamo					

SUMMARY

The figure shows the basic circuit diagram of the prototype with both stages operating in class B. For the calculation of the matching networks the complex input- and load impedances of the transistors have been calculated with the aid of a computer programme (ref. 1 and 2).

The average values are:

BFQ42: R_i = 6,7 ohm; X_i = -4,0 ohm; R_L = 41,8 ohm; C_L = -18,6pF

BLW29: R_i = 1,2 ohm, X_i = 0,4 ohm ; R_L = 5,5 ohm; C_L = -60,6pF

To reduce the costs and to simplify the lining-up the number of tuning elements has been reduced as much as possible. Furthermore the circuit contains a number of components to guarantee stability under mismatch conditions.

The amplifier is printed on a double sided copper clad epoxy glass-fibre board that is housed in a polycarbonate box.

A small intermediate heatsink is applied that has to be screwed to an external one.

Some typical results are: P_o = 14W; V_B = 13,5V.

f(MHz)	P_i(mW)	I_{tot}(A)	eff(%)	gain(dB)
146	63	1,76	58,9	23,5
153	99	1,64	63,2	21,5
160	79	1,85	56,1	22,5
167	66	1,73	59,9	23,3
174	74	1,82	57,0	22,8

The amplifier is stable for load mismatches up to a VSWR of 4,4 (any phase) combined with any input drive up to an output power of 14W.

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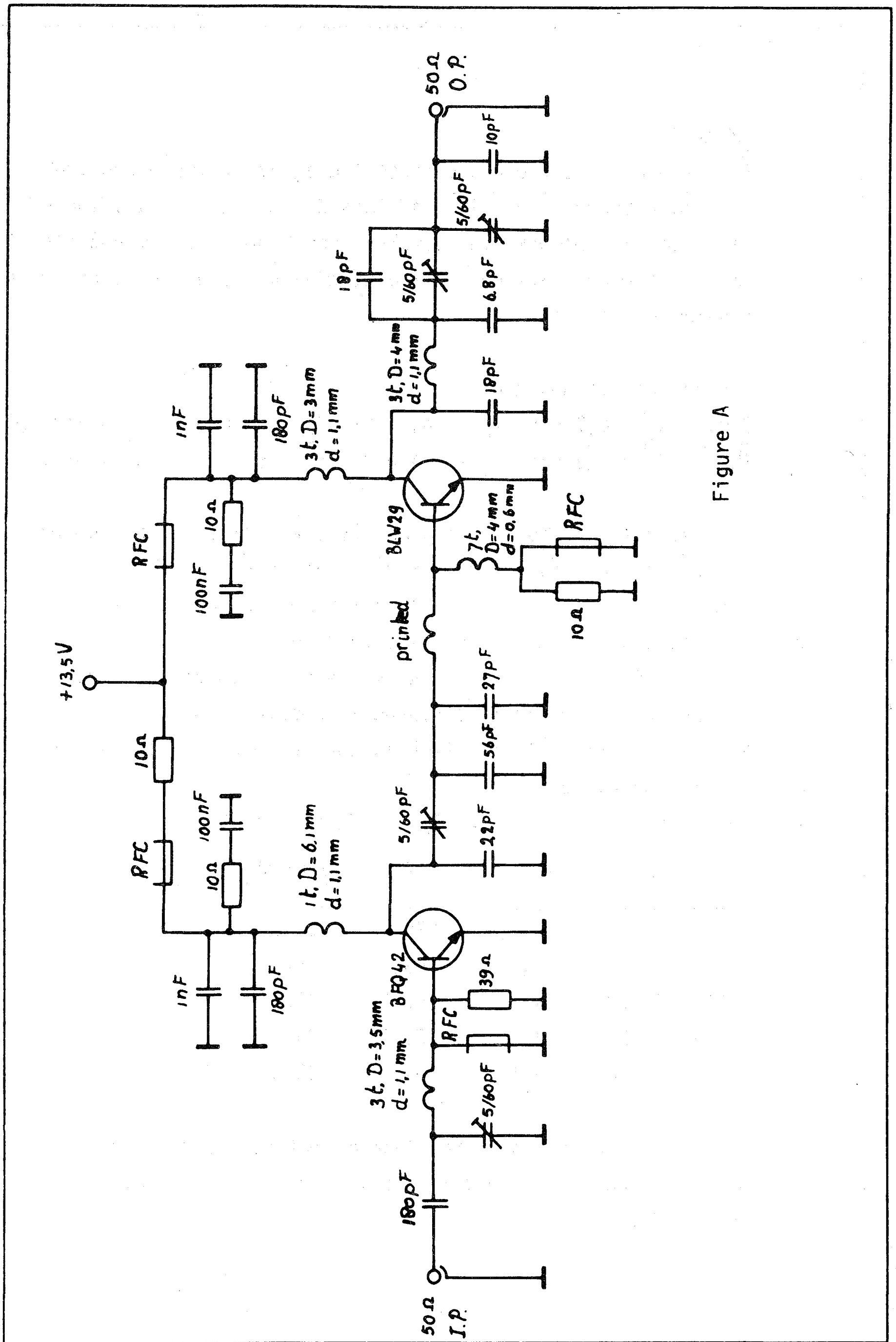

Figure A

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1. INTRODUCTION

The BFQ42 (dev. nr. 670 BLY) and BLW29 (dev. nr. 671 BLY) are VHF power transistors to be used in a two stage power amplifier for fixed or mobile communication equipment. The input power is less than 100mW for an output power of 14W.
The supply voltage is 13,5V.

This application is analogous to that of the BFQ43 and BLW31, which is described in C.A.B. report ECO 7701. The output power in that case was 28W.
The BFQ42 is mounted in a TO-39 metal envelope, with the collector connected to the case. The BLW29 is encapsulated in a ceramic envelope with 3/8 inch stud (SOT-120).

The amplifier is semi wide-band, the instantaneous bandwidth is about 2% and it is tuneable in the frequency range from 146-174 MHz. It is stable for an output VSWR of 4,4 (0-360^0) also with variation of the input drive power.

2. CIRCUIT DESCRIPTION

The construction of this two stage amplifier is the same as that of the 28W amplifier.

2.1. The output network of the BLW29

For the calculation of this network, we have to know the optimum load impedance of the BLW29. This has been calculated for P_o = 14W and V_b = 13,5V at a number of frequencies.
The average value of R_L and C_L is:

$$R_L = 5,5 \text{ ohm}$$
$$C_L = -60,6 \text{ pF} \rightarrow X_L = 16,4 \text{ ohm}$$

The frequency used in the calculations is 160 MHz.
This impedance is transformed in two steps to 50 ohm, see figure 1.

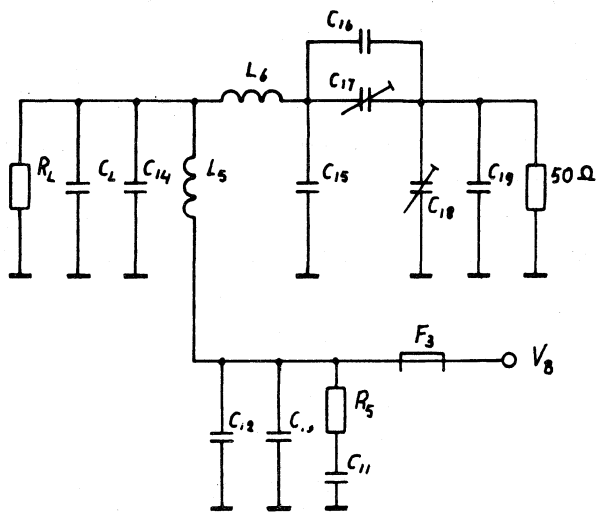

Figure 1

The first idea was to omit C_{15}, but during the experiments it became clear that it was better to split up the a.c. current after L_6.
L_5 is the r.f. choke. The reactance of which is about 5 to 7 times the collector load resistance.
The capacitors C_{12}, C_{13} and the series connection of R_5 and C_{11} together with the choke L_5 form a broadband collector impedance network that contributes to the stability when the external load is mismatched.
During the experiments with the 28W amplifier a capacitor from the collector to earth was added to improve the stability of the final stage. In the calculation of the output network of the 14W amplifier a capacitor of proportional value (C_{14}) was taken into account.

2.2. The interstage network BFQ42-BLW29

To calculate the interstage network we have to know the average output impedance of the BFQ42 and the average input impedance of the BLW29.

Both were calculated with the aid of a computer.

The results are: R_i = 1.2 ohm R_L = 41.8 ohm
X_i = 0,4 ohm C_L =-18.6pF → X_L = 53,5 ohm

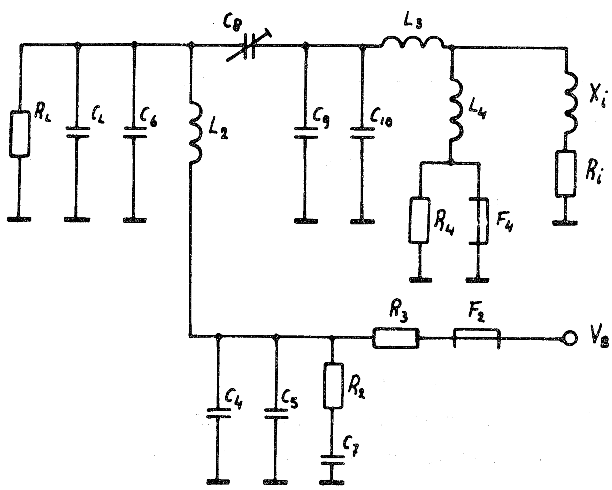

Figure 2

The collector of the BFQ42 is not directly connected to 13,5V for d.c. but a resistor is put in series. This is done because in this application the BFQ42 has to deliver only approx. 1W compared with a capability of 2W. In this way the stability of the driver stage is improved.

The collector choke L_2 forms a part of the interstage matching, i.e. the ratio of its reactance to the collector load resistance is much smaller than in the output network.

The capacitors C_4, C_5 and the series connection of R_2 and C_7 have the same function as similar components in the output network.

The capacitor C_6 is again added, to improve the stability of the driver stage.

The reactance of C_8 must be approx. -j22 ohm, which is realized with a trimmer of 60pF.
The inductance of L_3 is so small that it could be printed, see figures 5 and 8.

C_9 and C_{10} must be connected at 17mm from the transistor BLW29. Their leads have to be as short as possible.
The impedance of the combination L_4, R_4 and F_4 hardly influences the normal r.f. performance of the circuit. It consists of a carefully chosen combination that prevents parasitic oscillations under severe mismatch conditions.
The ferrite bead choke F_4 across R_4 is added because it is not advisable that the base current causes the transistor to operate in class C. This network gives a better suppression of parasitics than a single ferroxcube choke.

2.3. The input circuit of the BFQ42

The average input impedance of the BFQ42 is: R_i = 6,7 ohm
X_i =-4,0 ohm

This impedance has to be matched to 50 ohm, which can be done with one variable element.

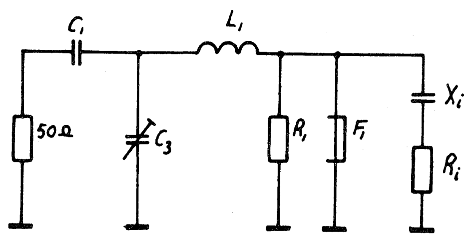

Figure 3

The calculation of this network is rather conventional.
The resistor R_1 and the choke F_1 form again a combination to suppress parasitic oscillations.

C_1 is a blocking capacitor of 180pF. This value has been chosen because of its series resonance around 160 MHz.

3. CONSTRUCTIONAL DETAILS

The printed circuit board is the same as used for the 28W version. The existing interruption in the base line of T_2 (L_3) must be short-circuited by means of copper foil (see figs. 5 and 8).

Further an interruption must be made in the supply line to the BFQ42, because a resistor (R_3) is connected in series. The board is of 1,5mm (1/16 in) double sided copper-clad epoxy glass-fibre with copper thickness of approx. 35 μ. Soldered rivets are inserted to make ground connections between the upper and lower sides of the board.

The intermediate aluminium heatsink is too small to handle the power dissipation. So it has to be screwed to a suitable one that can dissipate at least power levels of about 14W under non-mismatched conditions.

The p.c. board is screwed against the aluminium heatsink with the aid of two M2 screws and the stud of the BLW29.
The BFQ42 is mounted up-side down. Its case may not contact the heatsink electrically, so a boron-nitride disc is used in between to improve the thermal resistance. In addition some silicone-grease has been applied (see fig. 6).
A square hole in the p.c. board is made for the BLW29.
The advantage is that it is possible to make a direct contact from upper to lower sheet with wide copper strip (see figure 7).

4. MISMATCH TESTS AND SPURIOUS GENERATION

An important requirement on this amplifier is that it must be stable and has to survive mismatch tests applied at the output.

The stability tests have been done according to the following specifications:

. measuring frequencies: 146-153-160-167-174 MHz
. supply voltage V_b = 13,5V
. output power P_o = 14W
. drive level up to 100mW over 50 ohm
. VSWR (output) 1:4,4 (0-360^o)
. heatsink temperature approx. 20^{o}C

Tests have been made with the set-up shown in figure 9 in which the spectrum analyzer HP8558B appeared to be an indispensable instrument. A Rhode and Schwartz signal generator type SMLU has been used. The frequency counter is a Philips type PM6615. The forward and reflected power at the input are measured by a Rhode and Schwartz wattmeter and matching indicator, type NAU. The output power is attenuated 30dB by a Bird attenuator. The power is measured by a Hewlett-Packard powermeter type 435A. The mismatch tests were done with a normal 2dB power attenuator and a reactance unit, of which the internal diagram is shown in figure 10.

The current was measured with a Hewlett-Packard clip-on meter, model 428B.

5. MEASUREMENTS

At every test frequency, the interstage and output network were tuned for maximum output power and the input section for minimum reflection. The results are:

	P_o=14W			V_B=13,5V		
f(MHz)	P_i(mW)	P_r(mW)	I_{tot}(A)	I_{T1}(mA)	η%	gain(dB)
146	63	0	1,76	76	58,9	23,5
153	99	0	1,64	115	63,2	21,5
160	79	0	1,85	81	56,1	22,5
167	66	0	1,73	72	59,9	23,3
174	74	0	1,82	72	57,0	22,8

For an output power drop of 1W the instantaneous bandwidth is 2%. A 20% reduction of the battery voltage causes a maximum output power decrease of 37,1%.

For the German market the following adjustment may be of interest.

	P_o=7.5W			V_B=10.0V		
f(MHz)	P_i(mW)	P_r(mW)	I_{tot}(A)	I_{T1}(mA)	η%	gain(dB)
146	55	0	1,29	61	58,1	21,3
153	87	0	1,22	101	61,5	19,4
160	58	0	1,36	64	55,1	21,1
167	58	0	1,18	64	63,6	21,1
174	59	0	1,28	60	58,6	21,0

The amplifier is stable for load mismatches up to a VSWR of 4.4 (any phase) combined with any input drive up to an output power of 14W.

6. REFERENCES

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   Electronic applications bulletin, volume 31, number 3, 1972,
   p.p. 135-150, by A.H. Hilbers

2. Large-Signal behaviour of r.f. power transistors.
   Part 2, Computer programme.
   Electronic applications bulletin, volume 31, number 4, 1972,
   p.p. 218-233, by J. Mulder

3. Design of a semi-wide band power amplifier (146-174MHz) with
   BFQ43 and BLW31. C.A.B. report ECO 7701, by M.J. Köppen.

<u>Parts list</u> (Figs. 1-5)

R_1 = 39Ω, carbon ± 5%, CR25 style
R_2, R_4, R_5 = 10Ω, carbon ± 5%, CR25 style
R_3 = 10Ω, carbon ± 5%, PR37 style

C_1, C_5, C_{13} = 180pF, 100 V.d.c., ceramic
C_2 = omitted
C_6 = 2,2pF, 500 V.d.c., ceramic
C_3, C_8, C_{17}, C_{18} = 5/60pF, film dielectric trimmer (cat.nr. 2222 809 08003)
C_4, C_{12} = 1nF, 100 V.d.c., ceramic
C_7, C_{11} = 100nF, polyester ± 10%
C_9 = 56pF, 500 V.d.c., ceramic
C_{10} = 27pF, 500 V.d.c., ceramic
C_{14}, C_{16} = 18pF, 500 V.d.c., ceramic
C_{15} = 6,8pF, 500 V.d.c., ceramic
C_{19} = 10pF, 500 V.d.c., ceramic
F_1, F_2, F_3, F_4 = ferroxcube bead with 3 turns of 0,6mm Cuem. wire, cat.nr. of bead 4313 020 15172
L_1 = 3 turns, D_{int} = 3,5mm, d = 1,1mm Cuem. wire, spacing = 0,2mm, leads 2x5mm.
L_2 = 1 turn, D_{int} = 6,1mm, d = 1,1mm, Cuem. wire, leads 2x5mm.
L_3 = printed (see p.c. board)
L_4 = 7 turns, D_{int} = 4,0mm, d = 0,6mm Cuem. wire, closely wound, leads 2x5mm
L_5 = 3 turns, D_{int} = 3,0mm, d = 1,1mm, Cuem. wire, spacing = 0,2mm, leads 2x5mm.
L_6 = 3 turns, D_{int} = 4,0mm, d = 1,1mm, Cuem. wire, spacing = 0,2mm, leads 2x5mm.
T_1 = BFQ42 (670BLY)
T_2 = BLW29 (671BLY)

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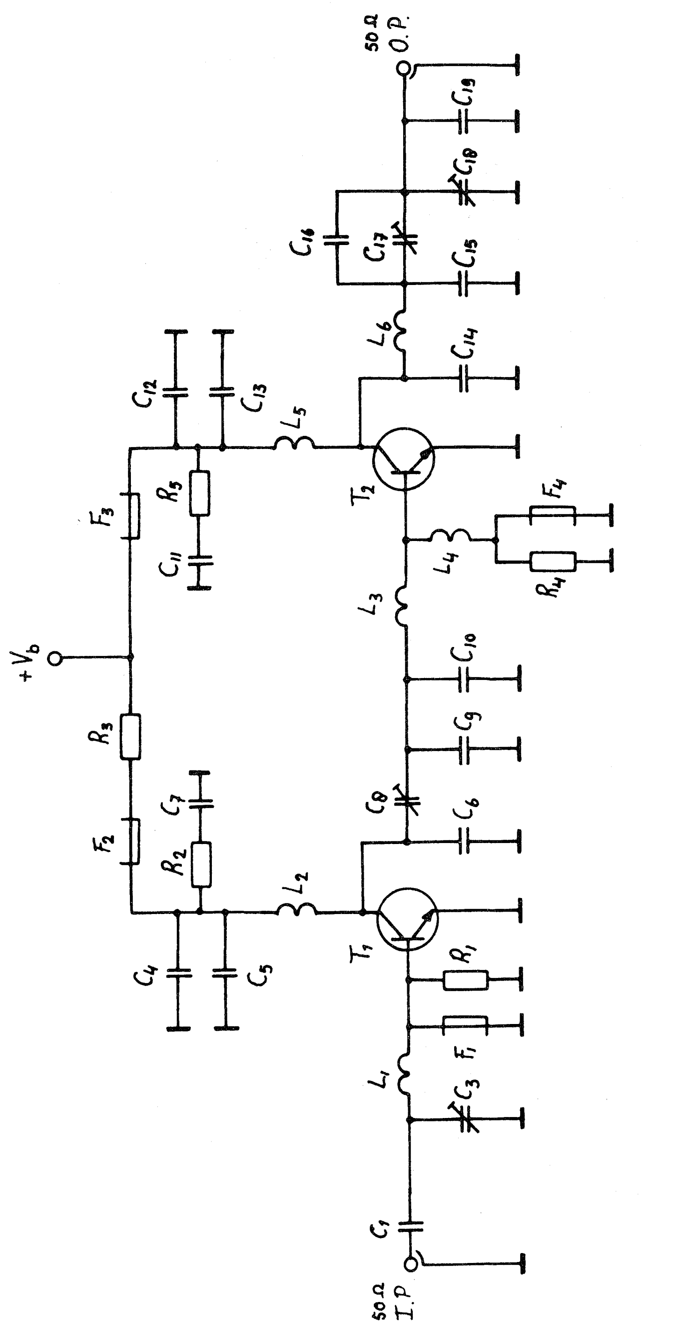

Figure 4.

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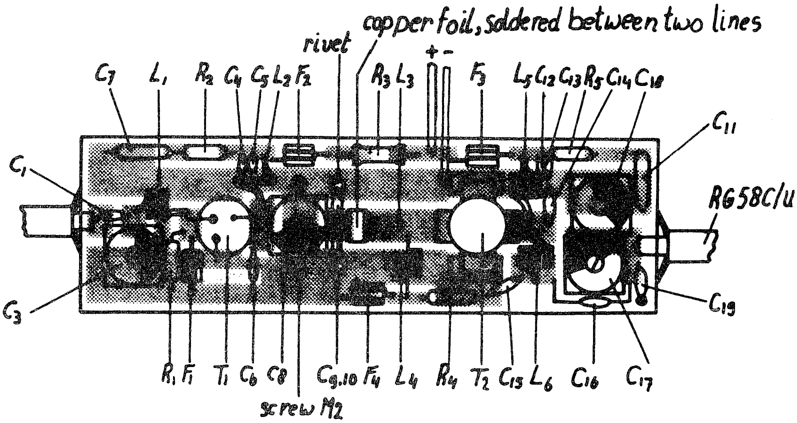

Figure 5

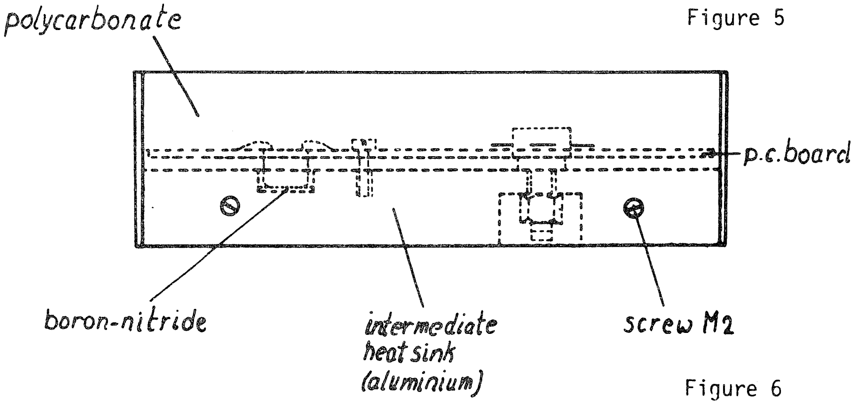

Figure 6

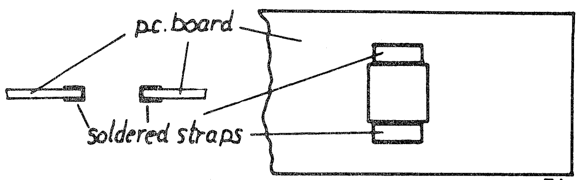

Figure 7

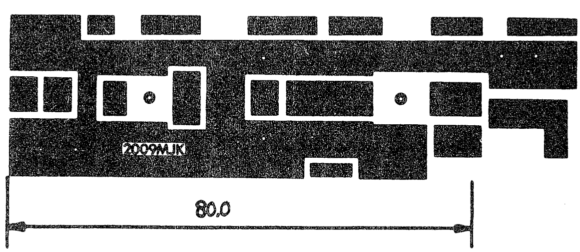

Figure 8

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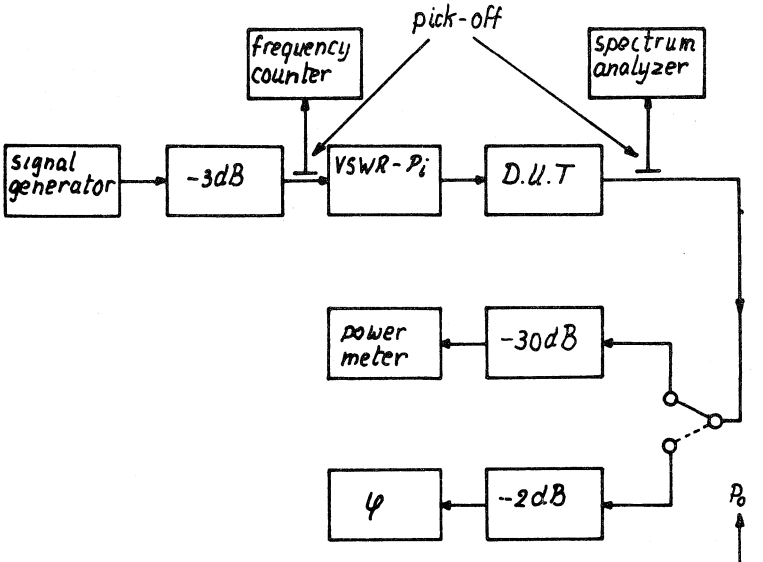
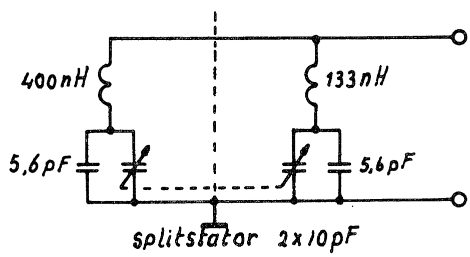


Figure 9

reactance (φ) unit



0 - 360°

VSWR ≥ 50

Figure 10

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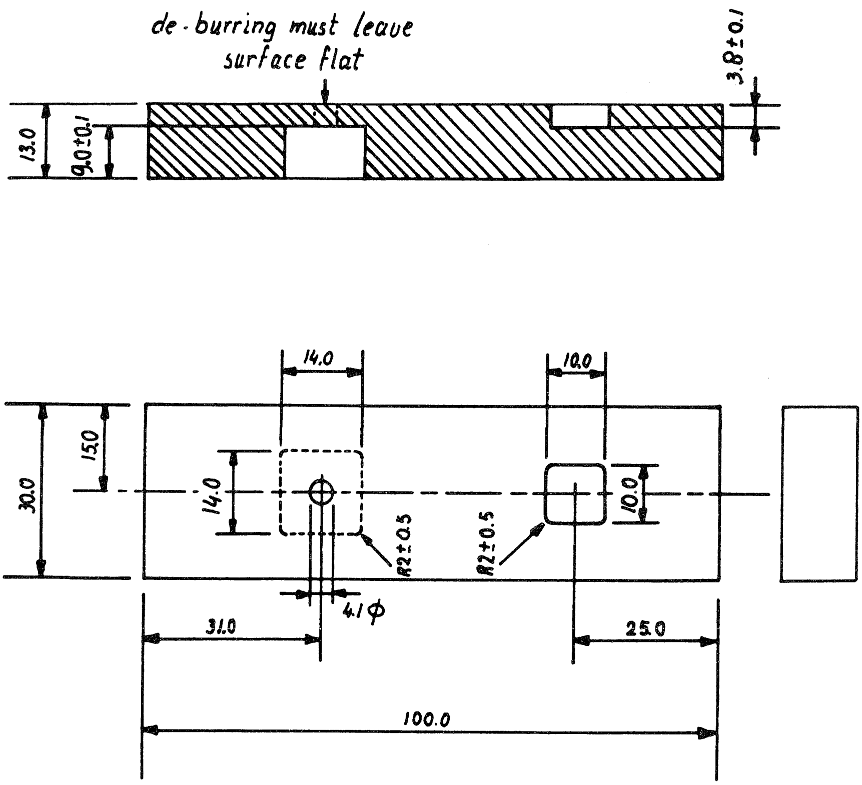

Dimensions in mm

Tolerances not specified : ±0.2 mm

Material : aluminium

HEAT SINK

Figure 11

Electronic
components
and materials

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laboratory report

number : EC08002 date : 1980-06-16

title : SURVEY OF TRANSISTORS FOR MILITARY COMMUNICATION TRANSMITTERS IN THE FREQUENCY BAND 25-80MHz.

author : A.H.HILBERS

Electronic components and materials

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laboratory report

central application laboratory CAB
eindhoven - the netherlands

number: ECO 8002 date: 16-06-1980

project: 6965 pages: A1 : R14

title SURVEY OF TRANSISTORS FOR MILITARY COMMUNICATION TRANSMITTERS IN THE FREQUENCY BAND 25-80MHz

author A.H. Hilbers

ABSTRACT

This paper presents a survey of transistor types suitable for application in military communication transmitters for the frequency band of 25-80MHz.
Some possible line-ups are discussed for portable, mobile and base station transmitters with different supply voltages and output power levels. Of the most important transistors power gain and impedance information has been given for wideband operation.

Appr. R.A. Pölzl

Advies Octrooi d.d. 1980-06-24	AV	X GV		B......		BL	
Opgave Mamo d.d. 1980-06-24	X AV	X GV	X SP	B...		BL	
Datum: 1980-06-17	Mamo						

1. INTRODUCTION

The Philips range of transmitting transistors counts nearly a 100 types which are generally specified at the maximum frequency of a civil communication or broadcast band. Except some typical information versus operating frequency none of the devices is explicitly specified for the military communication band of 25-80 MHz.
However a large percentage of our VHF and UHF transistors is well suited for application in this frequency band. It is the intention of this paper to elucidate this in more detail.

2. POSSIBLE LINE-UPS

In table I a list is shown of all the types suitable for a car battery voltage of 12-14V. Table II gives the types intended for supply voltages of 28V and higher. In both tables the power gain specified is the minimum one at 80 MHz in a tuned narrow-band amplifier.

To come to a series of transistor line-ups we have to consider some points:

a. In those cases where not only FM signals but also AM and/or SSB signals must be handled some linearity of the power amplifiers is required. This means that the output stage and for higher output powers also the driver stage must preferably be operated in class-AB push-pull. The remaining driver stages can then be class-A amplifiers with a suitable combination of voltage and current (negative) feedback.

b. The possibilities for impedance matching between the transistors and between the final stage transistors and the load are rather restricted. Conventional transformers can hardly be used because of their stray-inductance.
Chebyshev matching techniques according to Matthaei (Ref. 1.) are of restricted value because of the wide frequency band.

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Fig. 1 shows the maximum input VSWR in the frequency range of 25-80MHz for 2 and 3 section networks versus the impedance ratio. From this graph it can be seen that the maximum impedance ratio for a 2 section network is appr. 1,5 and for a 3 section network appr. 2 if we assume that the maximally acceptable VSWR is 1,25.

The most convenient way of impedance matching is probably the use of transmission line transformers with ferrite cores. These transformers have very low losses and a large power handling capability in a small volume. A disadvantage however is the fact that only some specific impedance ratios, viz. 1:1, 1:4, 1:9, etc. can be handled. (Ref. 2,3). This means that in a power amplifier of which a high collector efficiency is expected not every combination of supply voltage and output power can be chosen. If we restrict ourselves to impedance ratios of 1:1 and 1:4 (to reduce complexity) the load impedance of the final transistors is either 50 ohms or 12,5 ohms measured between the collectors. In this way it is possible to design transmitters for supply voltages of 26-28V with output powers of 20-25W or 80-100W. If the supply voltage is 13-14V this becomes 5-6W or 20-25W.

In driver stages where the efficiency is less important a load impedance below the optimum value can be chosen to preserve bandwidth. If a driver stage operates in class-A the allowable D.C. input power is appr. equal to the output power in class-B. The best efficiency obtainable in that case is between 30 and 35%. The design procedure for these stages is principally equal to that for application in the H.F. range (Ref. 4).

2.1 Portable transmitter with V_S = 7 - 8V

In Table III a series of line-ups for different supply voltages and output powers is given. The first design is intended for portable use. Fig. 2 shows the power gain and input impedance of the final stage transistor BFQ42.

At the output no transformation is required, only a balanced to unbalanced transformer. In some cases it can be useful to compensate the collector capacitance of the output transistors with an LC-network in the same way as described in Ref. 5 for conventional transformers.

2.2 Mobile transmitters with V_S = 13-14V

Two line-ups have been given. The first one with 2 transistors BLW80 in the final stage does not require impedance transformation at the output. In the second design where 2 transistors BLW29 are used an impedance transformation of 1:4 is required. Fig. 3 and 4 show power gain and input impedance of the mentioned transistors.

2.3 Mobile/base station transmitters with V_S = 26-28V.

The first design with 2 transistors BLY92C in the final stage requires no impedance transformation at the output.
In the second and third designs output transformers with a 1:4 impedance ratio are required. The difference between the second and third line-up is in the envelopes of the transistors.
The second design uses stud types and the third design flange types. Fig. 5 and 6 show power gain and input impedance of BLY92C and BLX39/BLW86.
In fig. 7 the same information is given for the BLY91C/BLV20 where used in a class-AB driver stage, loaded with 25 ohms per transistor. These types are then able to deliver min. 5W per transistor.

3. FINAL REMARKS

The power gain and impedance data given in this paper for some transistor types differs from the information given in the handbook. The reason for this is the different loading condition. In some transmitters stabilization of the supply voltage is applied. In a 28V system this could for instance mean that the supply voltage of the transistors is 22V. In such a case the same line-ups can in principle be used. One must however keep in mind that the output power varies with the square of the supply voltage if the output impedance transformation remains unchanged. To obtain a higher output power than indicated in the line-ups it is possible to combine 4 output stage modules by means of hybrid couplers as described in Ref. 2. and use this combination as a booster amplifier for the given line-up.
The information on the mentioned transistor types in this paper is rather restricted. More details can be found in the semiconductor handbook, part 4a. Additional information on other types for wideband operation like power gain and impedance levels can be supplied on request.

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Electronic components and materials

Type	V_C(V)	P_O(W)	G(dB)	Envelope
2N4427	13	1	18	TO-39 (1)
BFQ42	13	2	15	TO-39 (1)
BLW79	13	2	16	SOT-122
BFQ43	13	4	15	TO-39 (3)
BLW80	13	4	16	SOT-122
BLY87C	13	8	15	SOT-120
BLV10	13	8	14	SOT-123
BLW81	13	10	16	SOT-122
BLY88C	13	15	14	SOT-120
BLV11	13	15	14	SOT-123
BLW29	13	15	16	SOT-120
BLY89C	13	25	13	SOT-120
BLW87	13	25	13	SOT-123
BLW31	13	28	14,5	SOT-120
BLW60C	13	45	10,5	SOT-120
BLW85	13	45	9,5	SOT-123

Table I

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Type	V_C(V)	P_0(W)	G(dB)	Envelope
2N3866	28	1	19	TO-39 (1)
BLW89	28	2	20	SOT-122
BLW90	28	4	20	SOT-122
BLY91C	28	8	17	SOT-120
BLV20	28	8	17	SOT-123
BLW91	28	10	17	SOT-122
BLY92C	28	15	16	SOT-120
BLV21	28	15	16	SOT-123
BLX94C	28	20	16	SOT-122
BLY93C	28	25	16	SOT-120
BLW84	28	25	16	SOT-123
BLX39	28	45	13	SOT-120
BLW86	28	45	13	SOT-123
BLW75	28	60	13	SOT-105
BLV33	28	80	12	NO-207B
BLW78	28	100	10	SOT-121A
BLW95	50	150	8,5	SOT-121A
BLW96	50	200	7	SOT-121B

Table II

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Input power (mW)	1st stage	2nd stage	3rd stage	P_L (W)	V_{CE} (V)	Stud S Flange F
5	BFR96 1)	2 x BFQ42		1,5-2	7-8	-
15	2N4427 1)	2 x BLW80		5-6	13-14	S
50	BLW79 1)	2 x BLW29		20-25	13-14	S
50	BLW89 1)	2 x BLY92C		20-25	26-28	S
20	2N3866 1)	2 x BLY91C	2 x BLX39	80-100	26-28	S
20	2N3866 1)	2 x BLV20	2 x BLW86	80-100	26-28	F

Table III

1) Class-A operation

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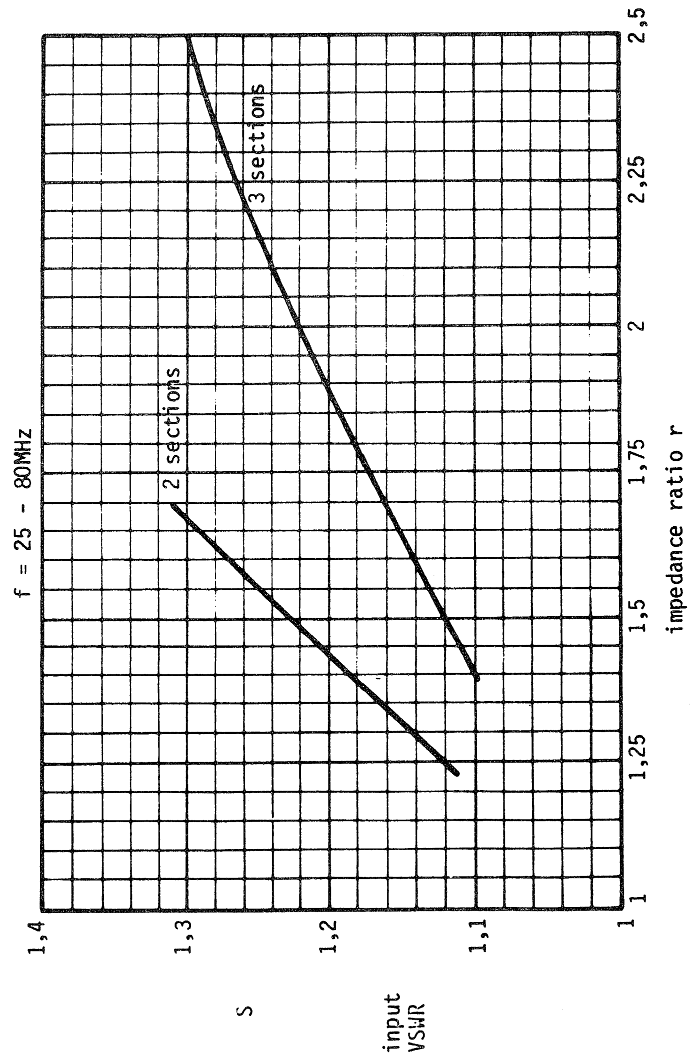

Fig.1

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8222 031 78361

BFQ42

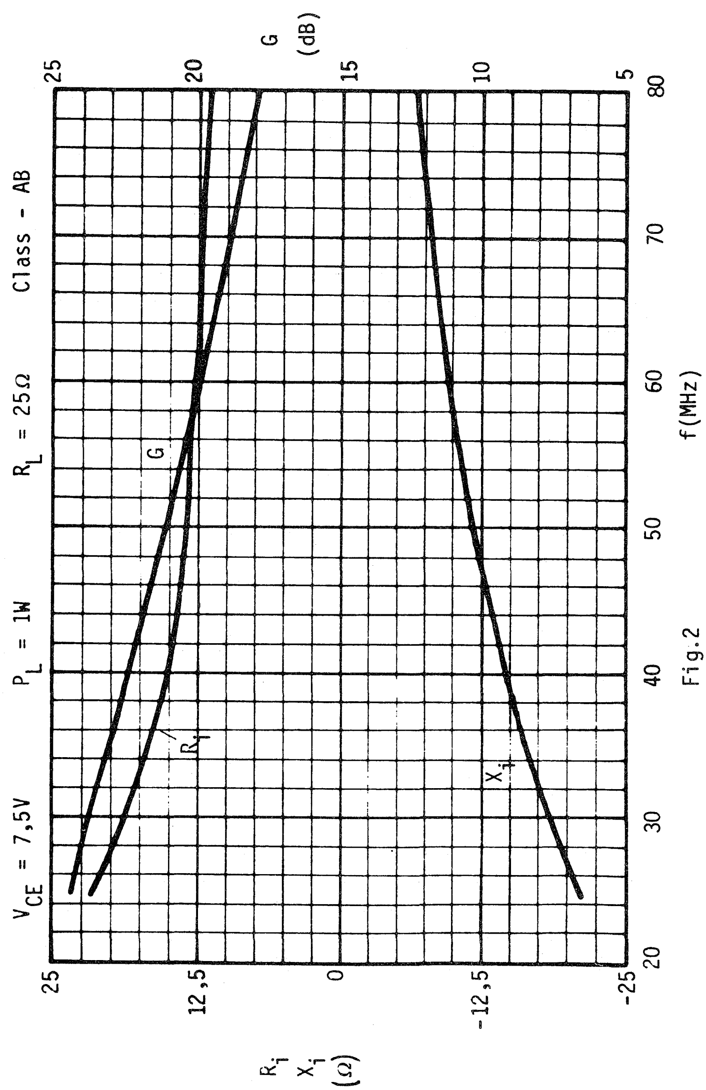

Fig.2

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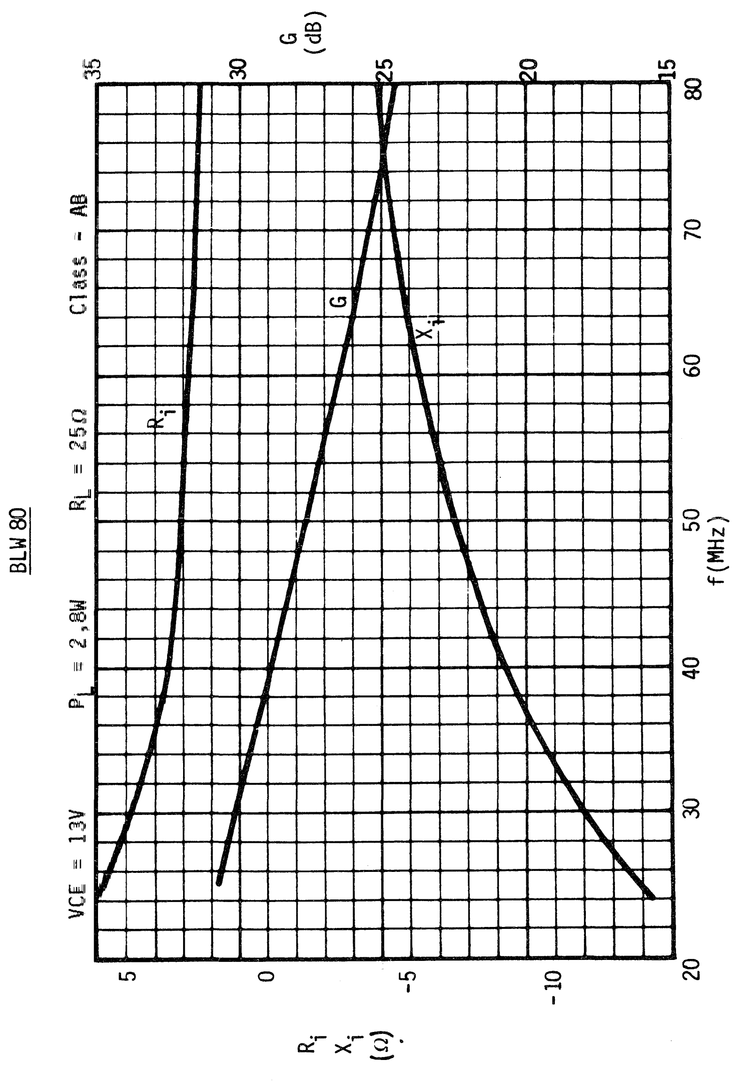

Fig. 3

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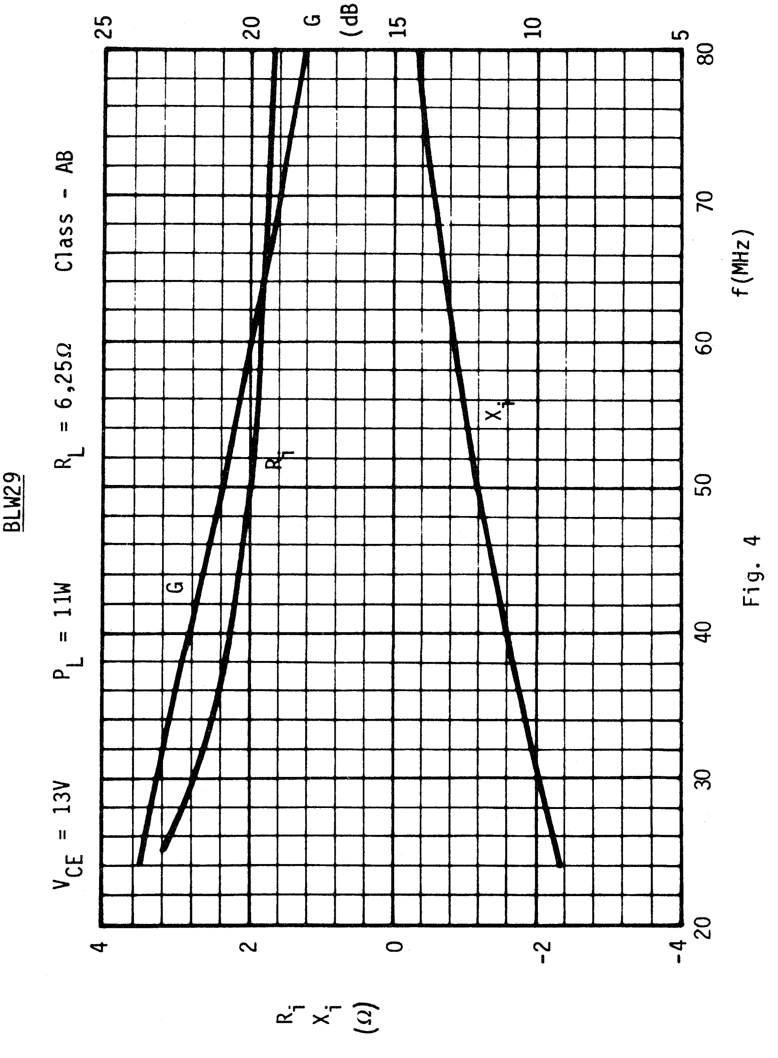

Fig. 4

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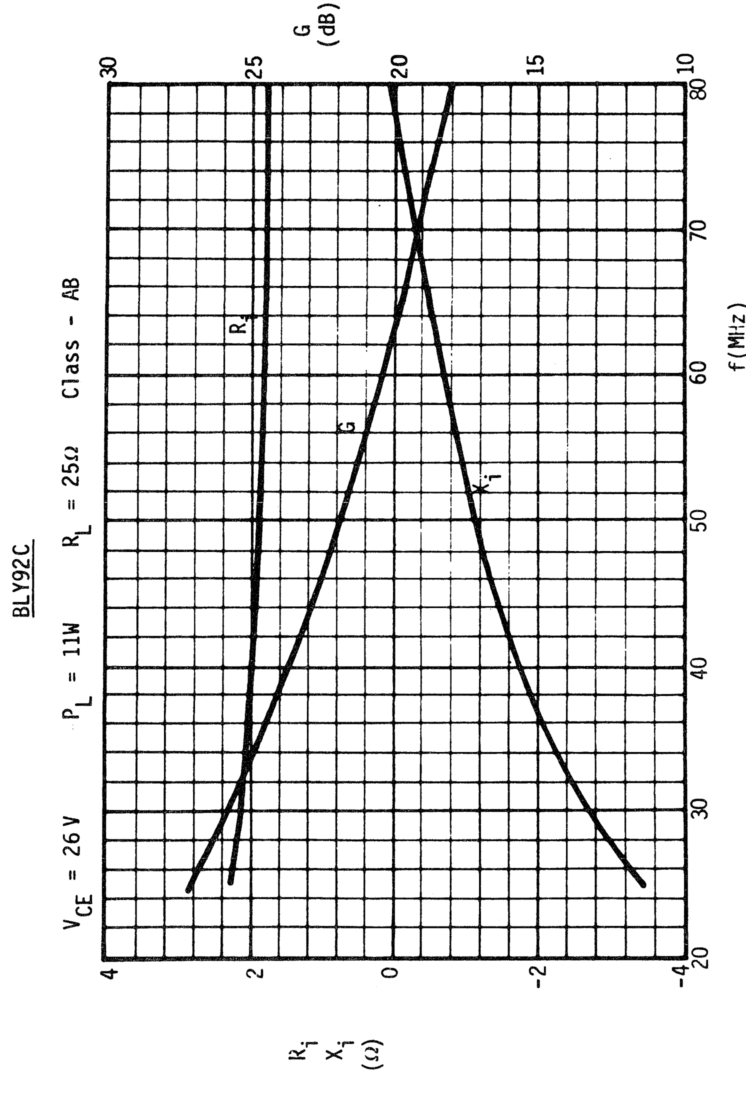

Fig. 5

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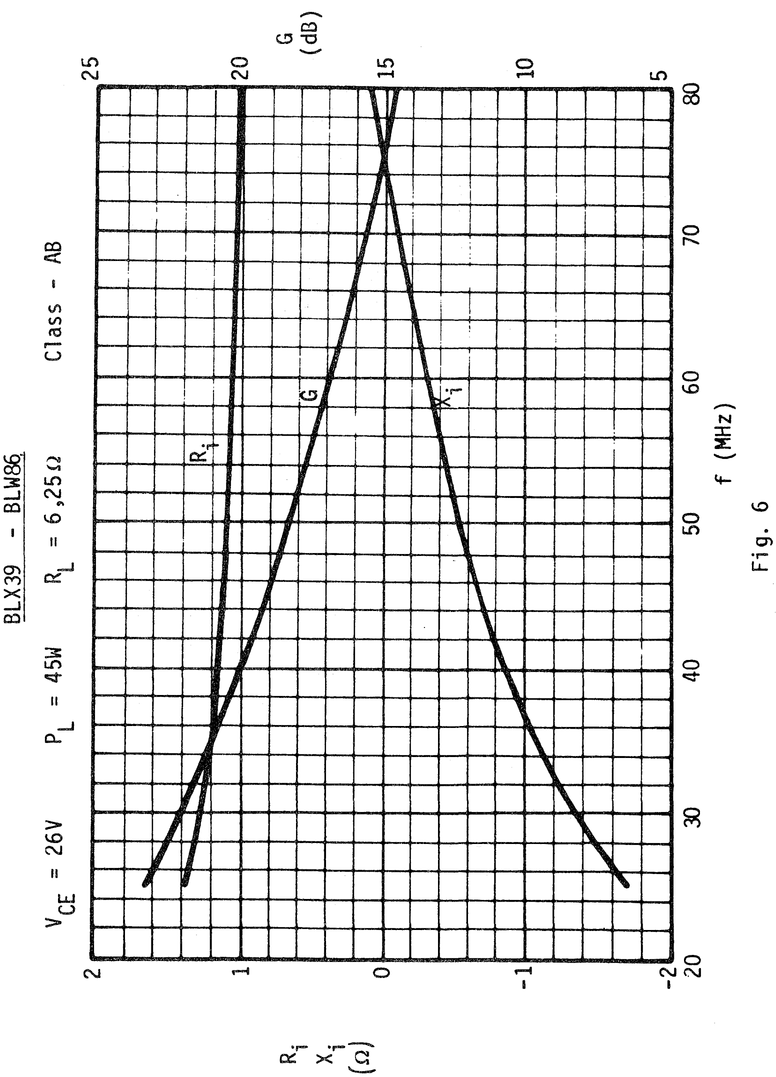

Fig. 6

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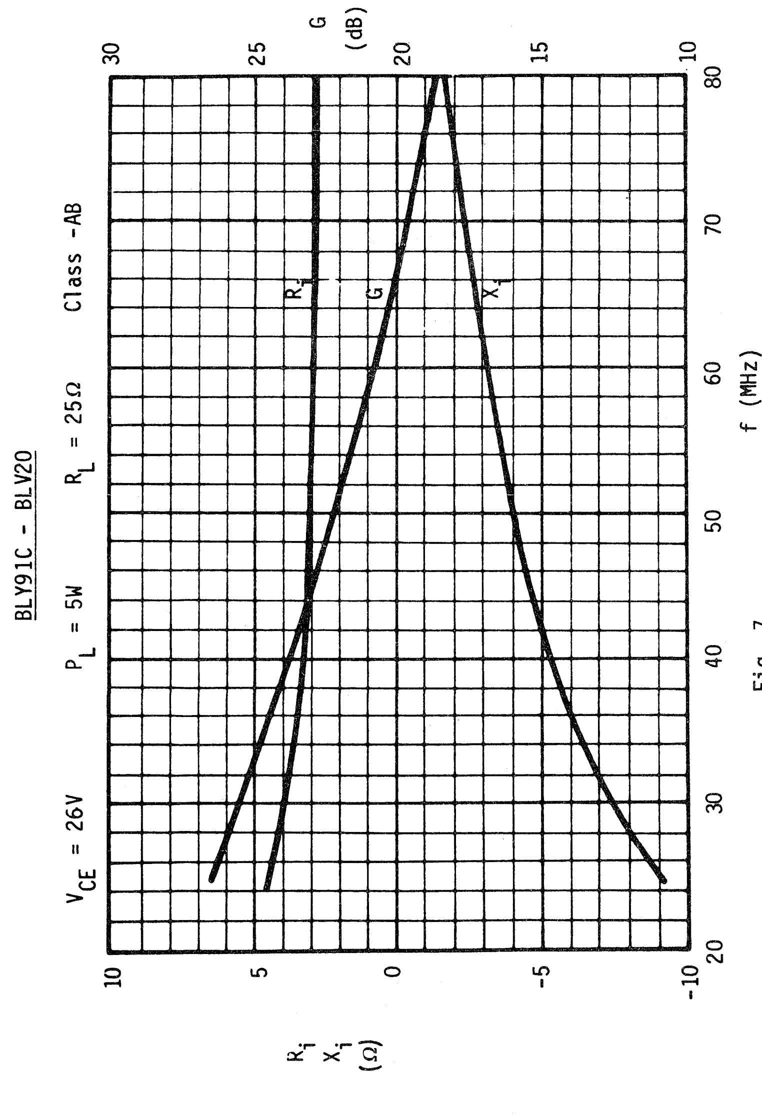
BLY91C - BLV20
V_{CE} = 26V
P_L = 5W
R_L = 25Ω
Class -AB
G (dB)
30
25
20
15
10
R_i X_i (Ω)
10
5
0
-5
-10
20
30
40
50
60
70
80
f (MHz)
R_i
G
X_i

Fig. 7

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eindhoven - the netherlands

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title SOME CONSIDERATIONS ON THE EFFICIENCY OF R.F.POWER TRANSISTORS IN THE DIFFERENT CLASSES OF OPERATION.

author A.H. Hilbers

summary

In this report considerations are given concerning the obtainable collector efficiencies in the different classes of operation of R.F.powertransistors. Also the frequency limitations are being considered.
As an example it can be mentioned that our 28V transistors are able to operate in class-E with an efficiency of 85% up to frequencies of 60 - 70 MHZ.

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1. INTRODUCTION

Some times we receive questions on the possibilities to improve the efficiency of R.F.power amplifiers. Below some considerations will be given for the different classes of operation of non-linear amplifiers.

2. CLASS-B OPERATION (V_{BE}= o)

For most of our transmitting transistors we publish the collector efficiency as measured in a class-B common-emitter narrow-band test circuit. Typically this efficiency is 65 - 70%.
This can be explained as follows:

a. the current efficiency of a class-B amplifier is:
   $\pi/4 * 100 = 78,5\%$
b. the loss of the output matching network is appr. 5% because it is designed for a loaded Q-factor of 10 whilst the unloaded Q-factor is appr. 200.
c. the remaining losses are D.C. and R.F.losses in the built-in emitter and collector resistances of the transistor. The former is necessary for a good D.C. SOAR and the latter for a high reverse second breakdown energy giving the device sufficient ruggedness against load mismatch.

However from this situation there are deviations in both directions; e.g. the BLW 89 has an efficiency of only 53%. This is caused by the tuning method of the narrow-band test circuit, because the circuit is tuned for maximum power gain. The corresponding load impedance of the transistor is some times strongly reactive. Phase angles of 40-45^{o} can occur which means that the collector efficiency must be multiplied with the cosine of this angle. If a different aligning method was applied, e.g. with a more resistive dummy load, the same transistor would show a higher efficiency at the cost of some power gain.

The other extreme is formed by transistors like the BLW 60 and BLY 90 having efficiencies of appr. 80%. This happens only at relatively high operating frequencies where the power gain is only 4-5dB. In such cases a substantial amount of the drive power is fed directly to the output circuit via the collector-base capacitance and the emitter lead inductance. This causes an artificially high collector efficiency. It is therefore better to compare transistors on the basis of oscillator efficiency being defined as:

$$\eta_o = \frac{P_o - P_i}{V_{ce}\, I_c}$$

In addition to the considerations above there is a slow decrease of efficiency when the operating frequency comes in the neighbourhood of 50% of the f_T of the transistor. This is caused by the increased capacitive current through the transistors collector and emitter resistances.

3. CLASS-C OPERATION

The collector current efficiency can be improved by reduction of the angle of current flow. This is achieved by a negative base-emitter D.C.voltage, e.g. caused by an external base resistor which is effective for D.C.

We do not recommend this method because a bias voltage in excess of 300 mV in combination with the increased drive voltage may cause continuous breakdown of the base-emitter diode leading to degradation of this diode, i.e. higher leakage current and reduced h'_{fe}.

On the other hand bias voltages below 300mV are not very effective.

4. CLASS-D OPERATION

An excellent description of this type of operation is given in Ref. 1. From this article some conclusions can be drawn:

a. push-pull operation is required with a very tight coupling between the two transistors. This can for instance be achieved by a complementary pair of transistors (NPN - PNP).

However pairs showing sufficient equality of D.C. and R.F. properties are not available.
As an alternative a matched pair of NPN-transistors can be used provided that a good combining transformer can be constructed which is only possible at frequencies up to appr. 30 MHz.

b. the maximum frequency of operation is further restricted by the switching times of the transistors. For output powers above 10W this is about 0,01 f_T.
This means for our modern transistors with U.H.F. diffusions a maximum of 5-10 MHz.

5. CLASS-E OPERATION

5.1 General considerations.

Design information on this class of operation can be found in Ref. 2 trough 5. The most important paper in this respect is Ref. 4, section III B, pp. 731-732, equations 3.20 through 3.29. The basic circuit diagram is reproduced in Fig. 1a en b. The reactance of RFC must be high compared with the load resistance. The series tuned circuit in Fig. 1b must have a sufficiently high loaded Q, e.g. greater than 5.
After some re-arrangement of the above mentioned equations we find that:

$$R_L = \frac{2V_s^2}{P_o(1+\frac{\pi^2}{4})} = 0{,}5768\,\frac{V_s^2}{P_o}$$

$$B = \omega_o C = \frac{2}{\pi(1+\frac{\pi^2}{4})R_L} = \frac{P_o}{\pi V_s^2}$$

$$X = \omega_o L = \frac{\pi}{8}(\frac{\pi^2}{2} - 2)R_L = \frac{\pi(\frac{\pi^2}{2} - 2)V_s^2}{4(1 + \frac{\pi^2}{4})P_o} = 0{,}6648\,\frac{V_s^2}{P_o}$$

An ideal transistor (having zero saturation resistance) will then show a collector efficiency of 100%.

Further we know from Ref. 4 that the collector peak voltage is 3,562 times the supply voltage and that the collector peak current is 2,862 times the collector D.C. current.

5.2 Practical example.

To show the possibilities of class-E operation we choose the BLV 25 which is able to produce 175 W of output power at a supply voltage of 28 V up to a frequency of 108 MHz. In a "normal" class-B amplifier this transistor shows an efficiency of 70-75%.

The guaranteed collector breakdown voltage of the BLV 25 is 65 V and the typical value 70 V. To prevent loss of efficiency by clipping of the collector voltage waveform we choose a supply voltage of 20 V.

The allowable collector peak current is 35 A. If we choose 30 A the drop of f_T is less than 20% compared with the point of maximum f_T .

Then the collector D.C. current becomes:

$$I_c = 30/2,862 = 10,5 \text{ A}$$

and the D.C. input power:

$$P_{dc} = 20 * 10,5 = 210 \text{ W.}$$

The saturation resistance is appr. 0,1 ohm and if we consider the collector current as a half sine wave the saturation loss can be approximated by:

$$P_{sat} = \frac{I_{cp}^2 * R_{sat}}{4} = \frac{30^2 * 0,1}{4} = 22,5 \text{ W}$$

So the transistor output power becomes:

$$P_o = P_{dc} - P_{sat} = 210 - 22,5 = 187,5\text{W}$$

As mentioned earlier the loss of the output matching network is 5%, so the output power of the amplifier will then be:

$$P_o' = 0,95 * P_o = 0,95 * 187,5 = 178\text{W.}$$

This gives us a collector efficiency of:

$$\eta_c = P_o' / P_{dc} = 178/210 * 100\% = 85 \%.$$

A question that still has to be answered is:
Up to what frequency can this performance be maintained?
The answer can be found by inspection of the formula for B in the previous section. We must keep in mind that C in this formula is the total capacitance from collector to ground, i.e. the sum of the transistors effective collector capacitance and a possible external capacitor. If we reduce the latter to zero and rearrange the formula we will find the maximum frequency of operation:

$$f_{max} = \frac{P_o}{2\pi^2 C_c V_s^2}$$

C_c is the effective collector capacitance which is 344 pF at a collector voltage of 20V, so:

$$f_{max} = 66 \text{ MHz}.$$

Unfortunately this is not sufficient to cover the F.M. broadcast band up to 108 MHz. However at least some of the advantage of class-E operation can be obtained by using the RFC to tune out the surplus of collector capacitance.
In this way a collector efficiency of 80% is probably possible.

6. FINAL CONSIDERATIONS.

An interesting question is whether the BLV 25 is a good or a bad transistor for class-E operation. Examination of many Philips and competition 28V transistors shows that it is a good average. f_{max} depends on the ratio Po/Cc which does not spread so much.
A much greater improvement is obtained by the application of 12V transistors. The quantity f_{max} rises then to the double value. However for a supply voltage of 12V there are no transistors available with the output power of the BLV 25.
If we would try to make such a device it would be a very impractical one, e.g. the optimum load resistance would be less than 0,5 ohm and the power gain 3dB less than that of the BLV 25, i.e. 7 - 8 dB.

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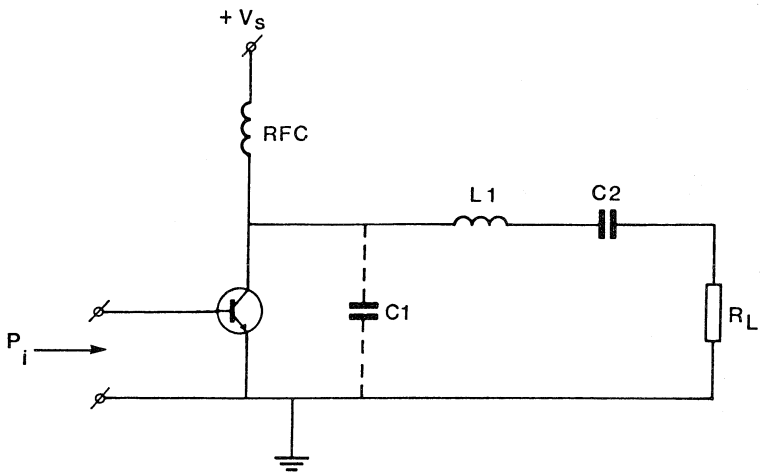

Fig. 1a

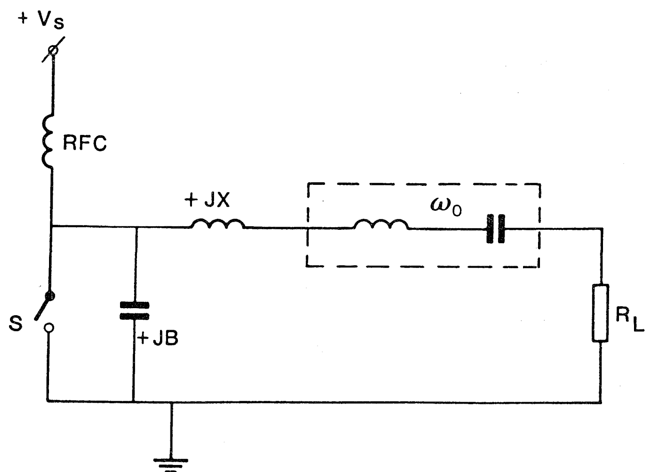

Fig. 1b

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NIJMEGEN – THE NETHERLANDS

REPORT No: NCO 8202	AUTHOR: H.G.van Hees
PROJECT No:	DATE: Nijmegen, 1982-05-17

TITLE

A WIDEBAND 300 WATT PUSH-PULL AMPLIFIER FOR THE FM BROADCAST BAND (87.5-108 MHz) WITH TWO TRANSISTORS BLV 25 AND A SUITABLE DRIVER AMPLIFIER

ABSTRACT

For the application in transmitters and transposers for the FM broadcast band (87.5-108MHz) a 300 Watt push-pull amplifier has been built with two transistors BLV 25. Also a suitable driver for this amplifier has been made with the transistor BLW 86.

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SUMMARY

For application in transmitters and transposers for the FM broadcast band (87.5-108MHz) a 300 Watt push-pull amplifier has been built with two transistors BLV 25. They operate with a class B setting, at a supply voltage of 28 Volt.
Moreover a single stage driver amplifier has been built with the transistor BLW 86, also operating in class B, at a supply voltage of 28 Volt.
The table shows the main properties of both amplifiers and of the combination of driver and final amplifier. The driver and final amplifier have been aligned at output powers of respectively 45 Watt and 300 Watt.

FM band 87.5-108MHz		Driver BLW 86 P_{out}=45W		Final 2xBLV 25 P_{out}=300W		Combination BLW 86-2x BLV 25 P_{out}=300W	
		min.	max.	min.	max.	min.	max.
Gain	dB	12	13	10.4	11	22.6	25
Input VSWR		1.2	1.3	1.45	1.70	1.1	1.85
Efficiency	%	69	72	70	71	63	66

The applied p.c. board is double copper clad epoxy fibre glass (ε_r=4.5), thickness 1/16 inch.

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The 2x BLV 25 amplifier has a heatsink with forced air cooling and a 10 mm copper plate working as a heatspreader.

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1. INTRODUCTION

The power transistor BLV 25 is intended for application in FM broadcast transmitters ans transposers. This device is encapsulated in a 6 leads flange envelope with a ½ inch ceramic cap. (SOT 119). In report ECO 8101 d.d. 11.11.1981 Mr.Hilbers gave some design considerations for a 300 Watt wideband push-pull amplifier for the FM broadcast band (87.5-108MHz) with two transistors BLV 25, operating in class B at a supply voltage Vs=28 Volt.
The practical solution for this theoretical design is described in the following chapters.
Besides a driver amplifier is described.
It is a single stage amplifier with the transistorBLW 86,designed for an output power of 45 Watt. The BLW 86 is encapsulated in a 3/8 inch 4 leads flange envelope with a ceramic cap.(SOT 123).
It is also operating in class B, with a supply voltage Vs=28 Volt.

2. 300 WATT PUSH-PULL AMPLIFIER WITH 2x BLV 25

2.1. General remarks

The schematic line-up of the amplifier is given in Fig.1.

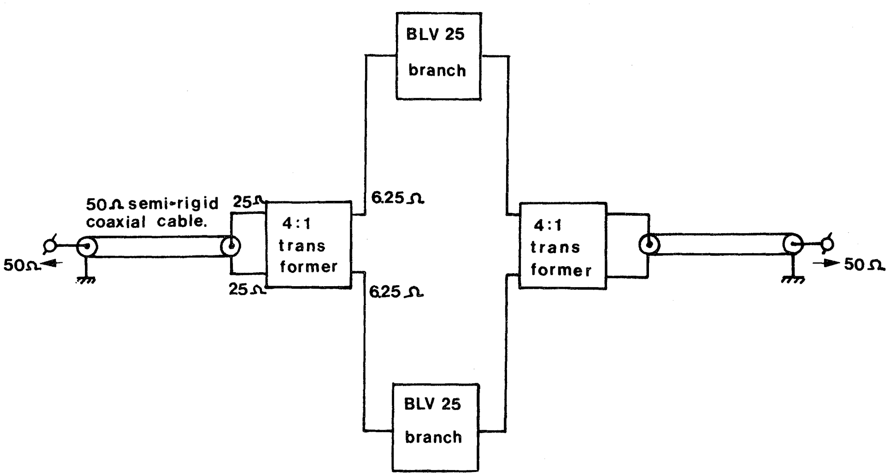

Fig.1

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Theoretical background and design details of this amplifier can be found in report ECO 8101 written by Mr.Hilbers.
The practical solution for the proposed electrical circuit is described here.
The amplifier has been designed on epoxy glass fibre print material ($\varepsilon_r = 4.5$), thickness 1/16 inch. Fig.2 shows the p.c. board and Fig.3 the lay-out of the ultimate amplifier.
For good contact between upper and lower side of the print, rivets have been used and at the edges copper straps have been soldered.
At the places where the emitters are grounded, contact has been established with the lower side of the print.
The printed circuit board and the transistors have been attached to a copper plate, thickness 10 mm, which functions as a heat-spreader. This plate has been screwed on a standard heatsink with forced air-cooling.
At an ambient temperature of 25^{o}C and the amplifier operating at 300 Watt output power level, the heatsink temperature is below 55^{o}C.

2.2. Alignment

The first alignment has been done on a small signal basis, starting with the output circuit. The BLV 25 transistors have been replaced by dummy loads, representing the complex conjugate of the optimum load impedance. The dummy consists of a 2.22Ω resistance and a 300pF capacitance.
To reduce parasitic inductance and to maintain the best possible symmetry we have used several components in parallel. These components have been soldered on an empty SOT 119 header.
The reflection versus frequency has been measured at the output terminal and minimized by adjusting the capacitors C_{16}, C_{14}, C_{15} and C_9. Fig.4 shows the return losses; VSWR remains below 1.13.

The alignment of the input network has been done with the transistors in the circuit and the supply voltage and load connected.

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We have started the alignment with the transistors in a class A setting (I_c=1.7A and V_{CE}= 25V). The reflections versus frequency then have been minimized at a small signal basis.
After that the transistors have been set in class B and the amplifier has been realigned at an output power level of 300 Watt. The required circuit modifications were in general rather small. Fig. 5 shows the final circuit. The resistances R_2 and R_3 are necessary to prevent parallel oscillations. The inductance of these resistors is very important (see parts list).
In spite of the dummy adjustment of the output circuit we had to lower the capacitance value of C_9, in order to improve the collector efficiency η of the amplifier. Besides we have applied three capacitors in parallel because of the very high reactive loading at that point.

2.3. Performance

The amplifier has been aligned at an output power of 300 Watt. Fig.6 and 7 respectively show gain and input VSWR and collector efficiency η as a function of frequency at 300 Watt output power. In Fig.8 is the efficiency versus output power and in Fig.9 the gain versus output power, both measured at 108 MHz.

3. BLW 86 DRIVER AMPLIFIER

3.1. Design of the amplifier

The wanted drivepower for the 2xBLV 25 amplifier, described in section 2, amounts to about 30 Watt. The input VSWR of this final amplifier varies between 1.45 and 1.7 (see Fig.6) thus the load impedance of the driver amplifier differs from 50Ω and varies as a function of frequency. Because we could not predict the effect of this on the performance of the driver, we have built in some reserve in output power and designed a 45 Watt driver. It is a one stage class B amplifier with the transistor BLW 86.

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Table 1 shows some properties of the BLW 86 in the frequency range from 87.5 to 108MHz, valid for class B operation and an output power of 45 Watt. Table 1

Freq. (MHz)	Gain (dB)	Input impedance (Ω)	load impedance (Ω)
87.5	13.61	0.76 -j 0.00	7.65 + j 3.28
89.8	13.40	0.76 +j 0.04	7.56 + j 3.32
92.2	13.18	0.76 +j 0.08	7.48 + j 3.36
94.7	12.96	0.76 +j 0.12	7.39 + j 3.40
97.2	12.75	0.75 +j 0.16	7.32 + j 3.47
99.8	12.53	0.75 +j 0.20	7.23 + j 3.51
102.5	12.31	0.75 +j 0.24	7.13 + j 3.54
105.2	12.10	0.75 +j 0.28	7.05 + j 3.60
108	11.89	0.75 +j 0.32	6.95 + j 3.63

The input impedance has to be matched to the 50Ω source impedance to obtain a good input VSWR and the 50Ω load impedance has to be transformed into the optimum load impedance, which is given in the table.
This has been done with Chebychev low-pass L-C filter techniques (Ref.1). The amplifier has been designed on double clad epoxy glass fibre print material (ε_r=4.5), thickness 1/16 inch. Fig.10 shows the p.c. board and the lay-out of the amplifier. Again rivets and straps have been applied and the emitter soldering places have been connected to the lower side of the print.

3.2. Alignment

The alignment has been carried out in the same way as described in chapter 2.2. (page 5). The optimal load impedance given in table 1 suggested a dummy load of 10Ω resistance in parellel with a 91pF capacitance. Alignment with this dummy resulted in an collector efficiency of about 60%.
Later on we found out that this dummy capacitance had to be decreased to 56pF to reach an efficiency of about 70%.
Fig.11 shows the VSWR at the output terminal measured with the 10Ω //56pF dummy inserted.

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The input circuit has been aligned with the transistor in the circuit and the load and the supply voltage connected. Again we have started the alignment with the transistor in a class A setting (Ic=1A and V_{CE}=25V). The small signal input VSWR has been minimized.

After that the transistor has been set in class B and the amplifier has been realigned at an output power of 45 Watt. Fig.12 shows the ultimate circuit. The collector d.c. biasing coil L_8 plays an active role in the transforming precedure.

3.3. Performance

Fig.13 shows gain and input VSWR as a function of frequency. Gain is 12.5dB $\pm$ 0.5dB and VSWR remains below 1.3:1 through the band.
In Fig.14 one can see that the collector efficiency η is better than 69%. The measurements have been carried out at an output power level of 45 Watt.
Fig.15 and 16 respectively show collector efficiency and amplifier gain versus output power at 108MHz. The amplifier has only been aligned at 45 Watt output power level.

4. COMBINATION OF DRIVER AND FINAL AMPLIFIER

Fig.17 shows tha gain and input VSWR of the combination of driver and final amplifier at 300 Watt output power. We have done this in order to get an idea of the influence of the fluctuating input VSWR of the final amplifier on the performance of the driver amplifier. The efficiency of the combination is better than 63%, as one can see in Fig.18.
No additional alignment has taken place. The required input power for 300 Watt output power amounts to less than 1.7 Watt.

5. CONCLUSION

We have built a 300 Watt push-pull amplifier with two transistors BLV 25. To obtain drive power we have also built a single stage driver amplifier with the transistor BLW 86.

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The properties of these amplifiers and of their combination are in the table below.

FM band 87.5-108MHz		Driver BLW 86 P_{out}=45W		Final 2xBLV 25 P_{out}=300W		Combination BLW 86-2xBLV 25 P_{out}=300W	
		min	max	min	max	min	max
Gain	dB	12	13	10.4	11	22.6	25
Input VSWR		1.2	1.3	1.45	1.70	1.1	1.85
Efficiency η	%	69	72	70	71	63	66

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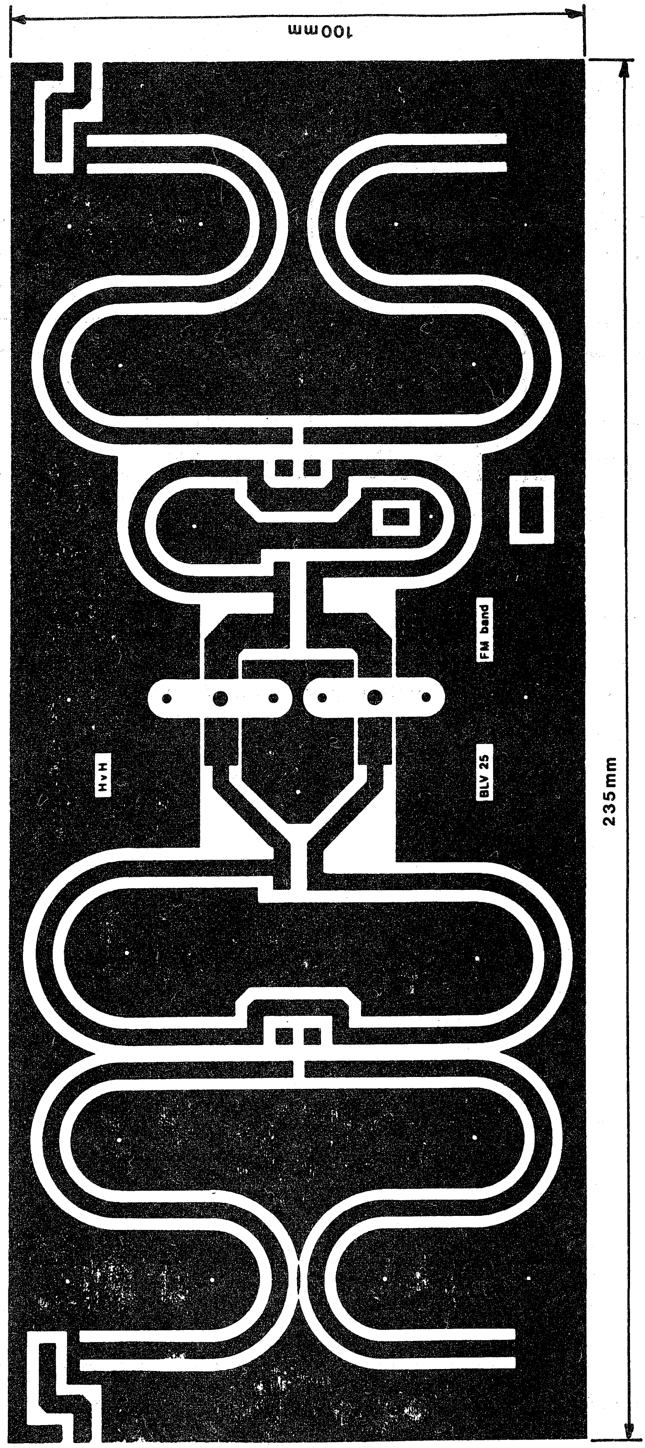
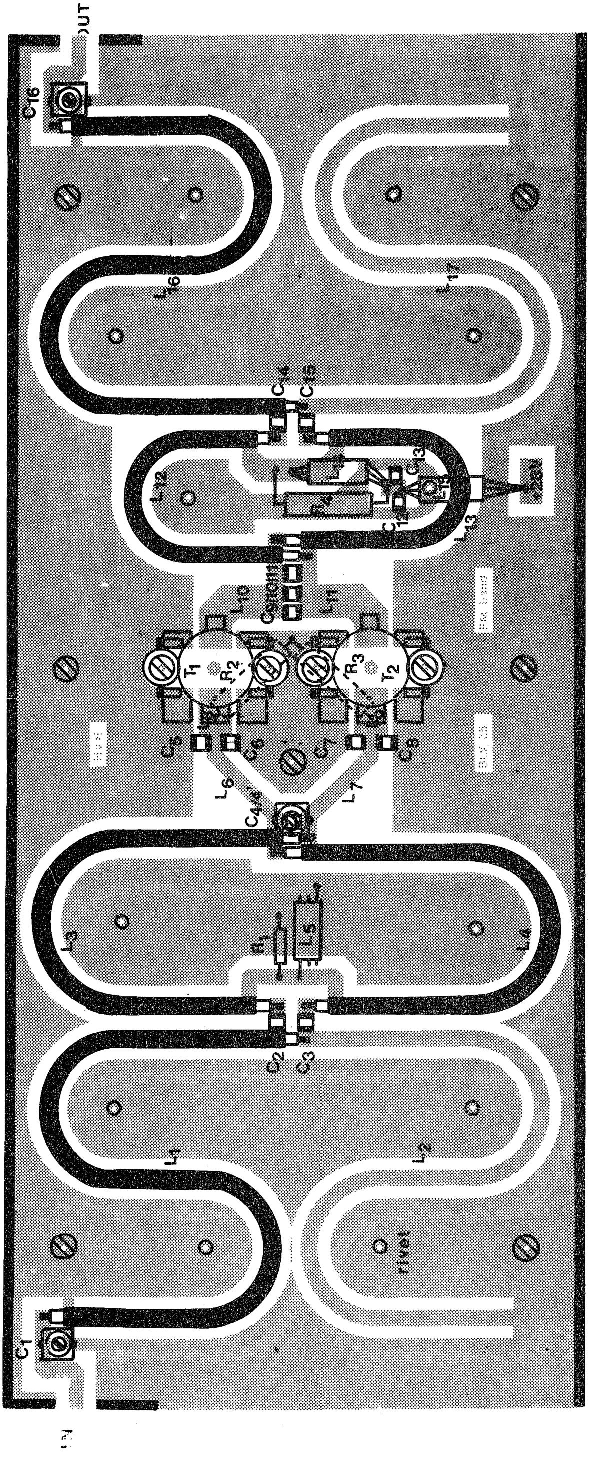
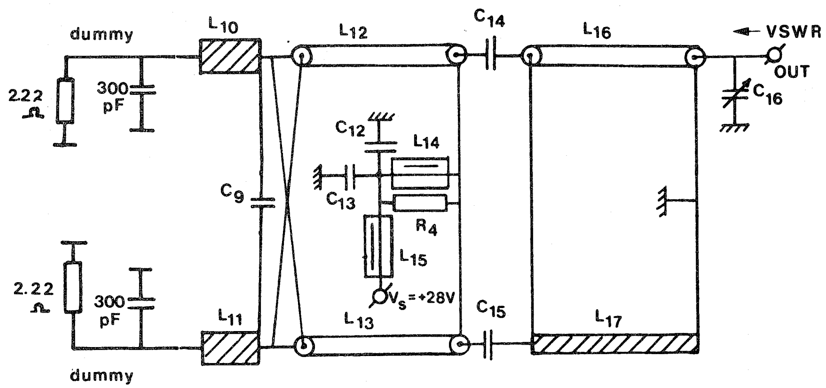


Printed circuit board Fig.2



Amplifier lay-out Fig.3



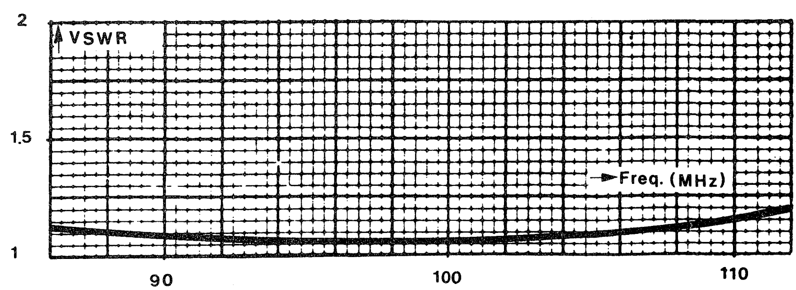


Fig.4 Alignment output circuit

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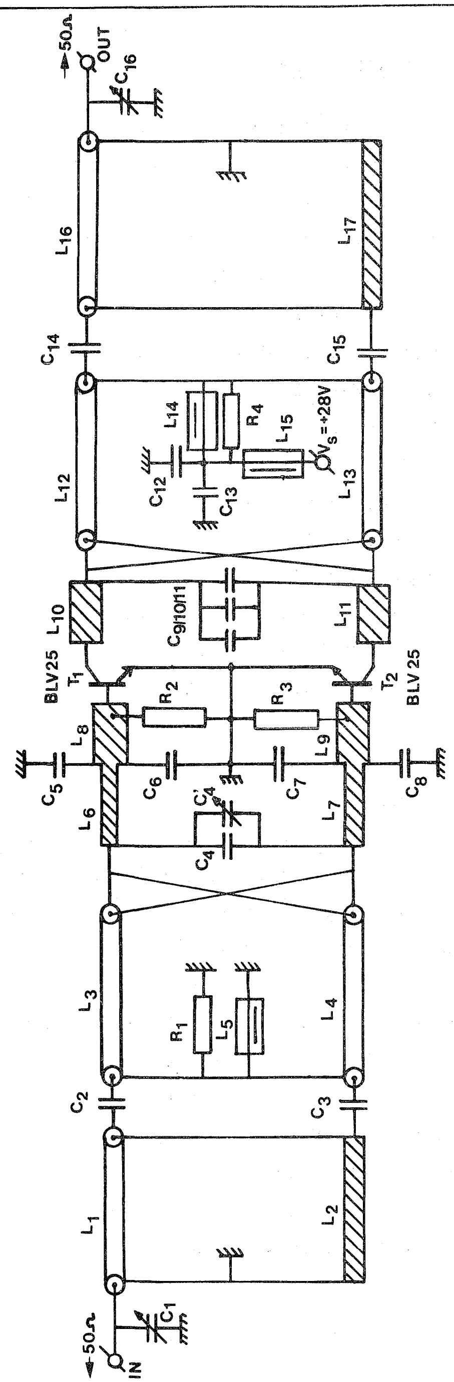

Fig.5 Amplifier circuit diagram

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PARTS LIST

R_1=12.1Ω metal film Philips MR 25 cat no.2322 151 71219

R_2=R_3=4.99Ω metal film Philips MR 52 cat no. 2322 153 54998

R_4=12.1Ω metal film Philips MR 52 cat no. 2322 153 51219

C_1=C'_4=C_{16}=2-18pF film dielectric trimmer Philips cat no.2222 809 05003

C_2=C_3=200pF chip ATC 100B-201-K-Px-300

C_4=300pF chip ATC 100B-301-K-Px-200

C_5=C_6=C_7=C_8=680pF chip in parallel with 150pF chip.
ATC 100B-681-K-Px-50
ATC 100B-151-J-Px-300

C_9=43 pF chip ATC 100B-430-J-Px-500

C_{10}=68pF chip ATC 100B-680-J-Px-500

C_{11}=82pF chip ATC 100B-820-J-Px-500

C_{12}=2k7 chip Philips NPO size 1210 cat no 2222 852 13272

C_{13}=100k chip Philips X7R size 1812 cat no 2222 852 48104

C_{14}=C_{15}=100pF chip ATC 100B-101-J-Px-500

L_1=50Ω semi-rigid coaxial cable, d=2.2 mm, l= 144 mm:
soldered on a 50Ω stripline, w= 2.8mm

L_2=50Ω stripline, w=2.8mm, l=144 mm

L_3=L_4=25Ω semi-rigid coaxial cable, d=3.5mm, l=96mm;
soldered on a 50Ω stripline, w=2.8 mm.

L_5=FXC 3B r.f. choke Philips cat.no 4312 020 36642

L_6=L_7=50Ω stripline w= 2.8mm, l= 18.1 mm

L_8=L_9=30Ω stripline w= 6 mm, l= 4.8mm

L_{10}=l_{11}=30Ω stripline w= 6mm, l= 14.1mm

L_{12}=L_{13}=25Ω semi-rigid coacial cable, d=3.5mm,l=60.3mm
soldered on 50Ω striplines, w= 2.8mm

L_{14}=L_{15}=FXC 3B beads, Philips cat no. 4312 020 31500
wound with 6 leads in parellel

L_{16}=50Ω semi-rigid coacial cable, d= 3.5mm, l= 139.6 mm;
soldered on 50Ω stripline, w= 2.8mm

L_{17}=50Ω stripline, w=2.8 mm, l= 139.6 mm.

T_1=T_2= BLV 25

p.c. board material: 1/16" epoxy fibre-glass, ε_r=4.5

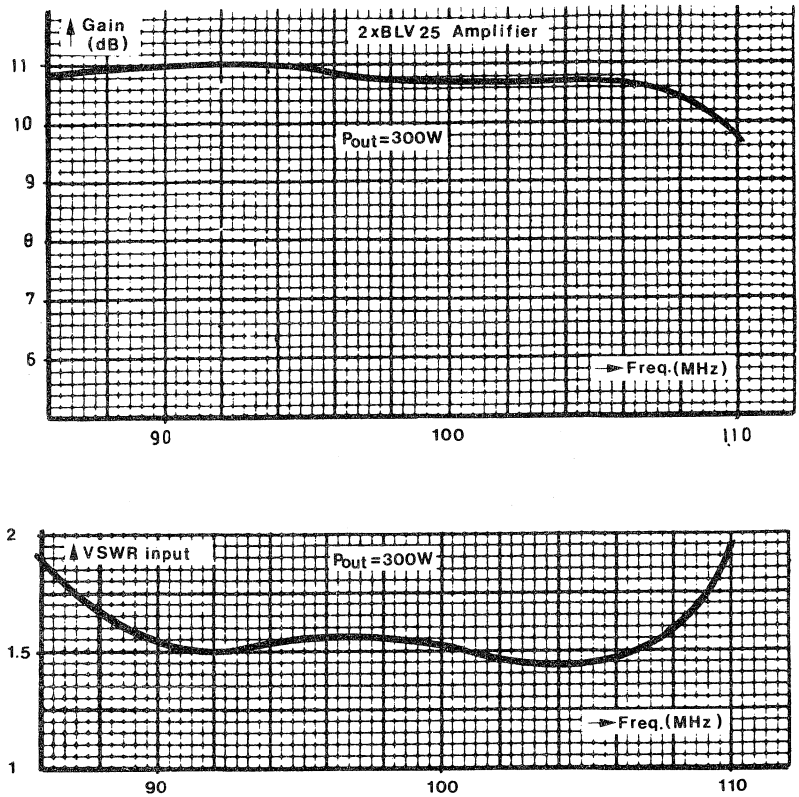

Fig.6 gain and input VSWR versus frequency

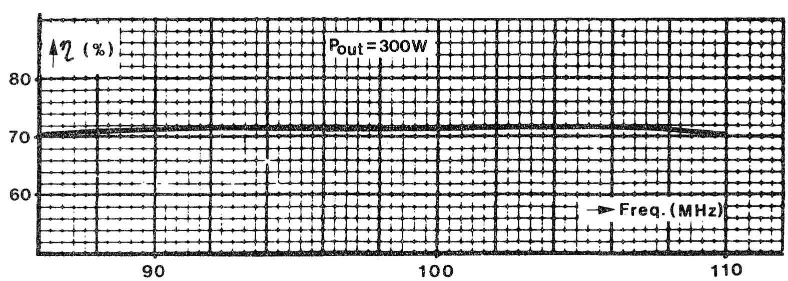

Fig.7 collector efficiency η versus frequency

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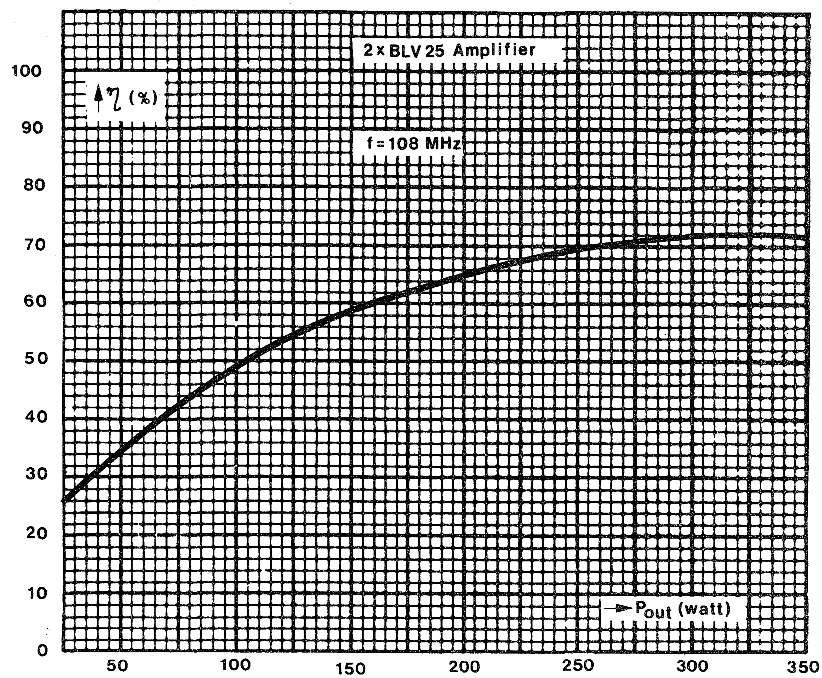

Fig.8 efficiency η versus output power

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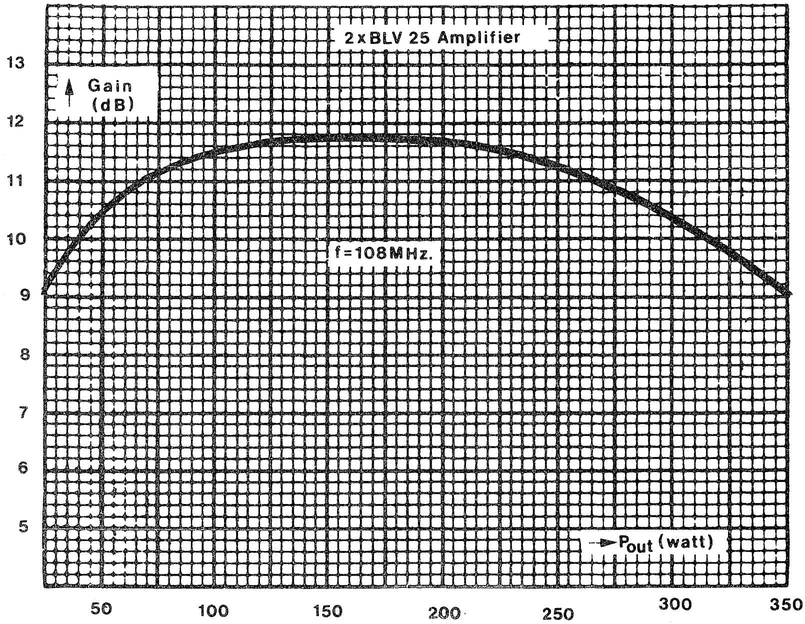

Fig.9 gain versus output power

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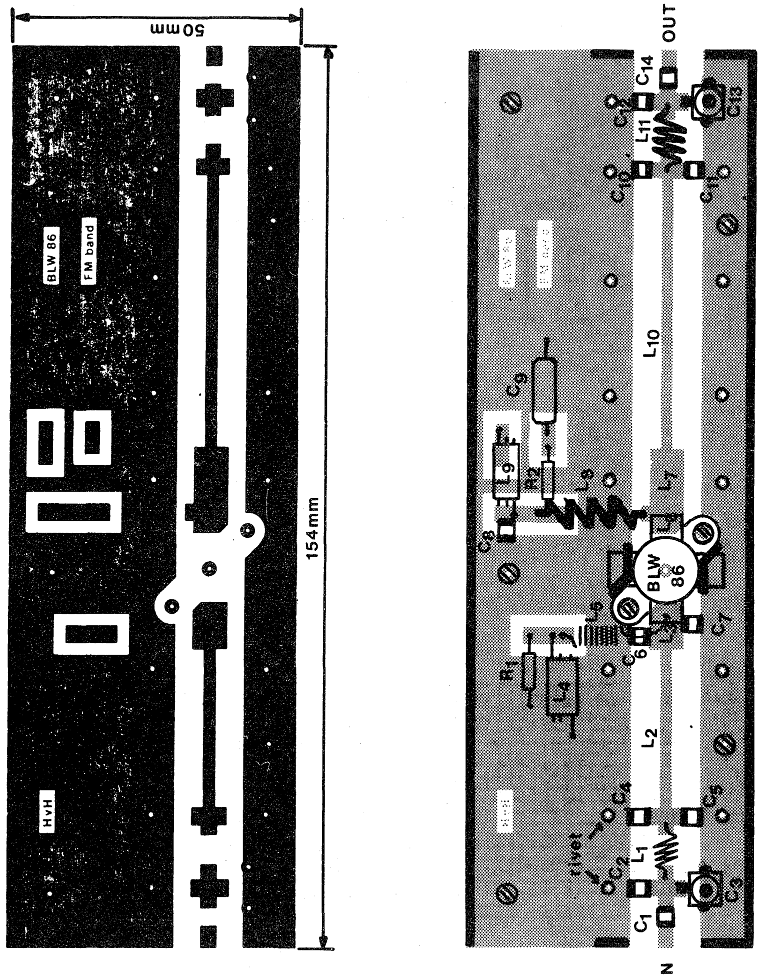

Fig.10 printed circuit board and lay-out

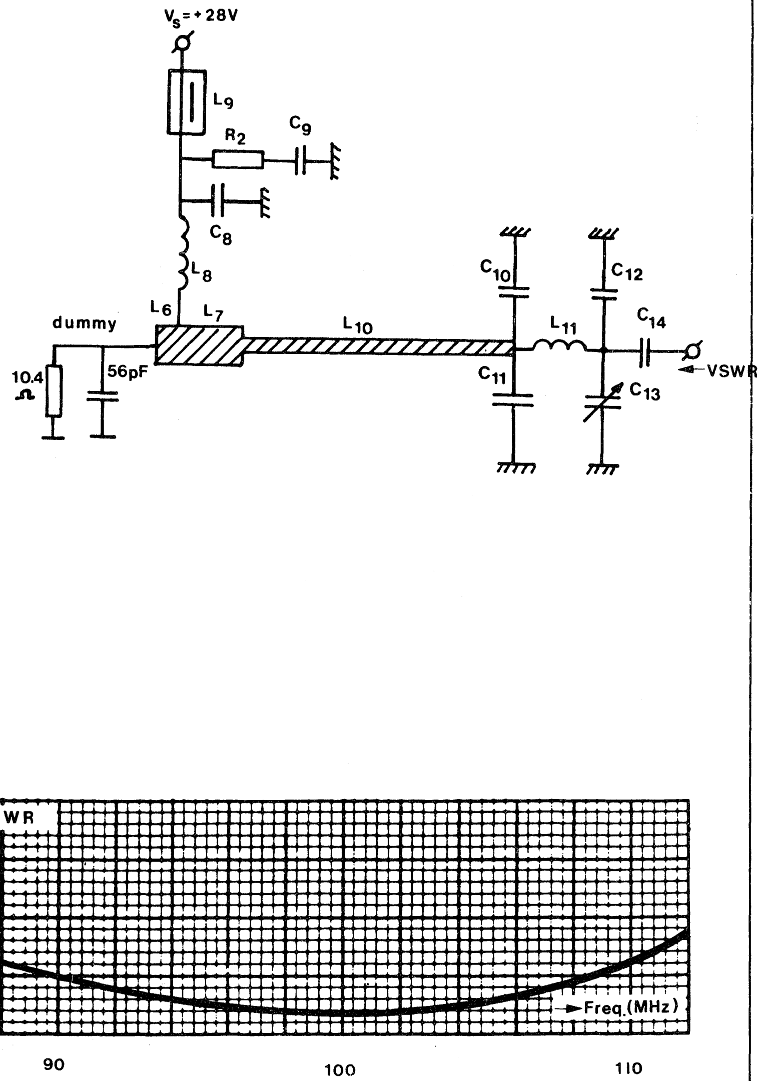

Fig.11 alignment output circuit

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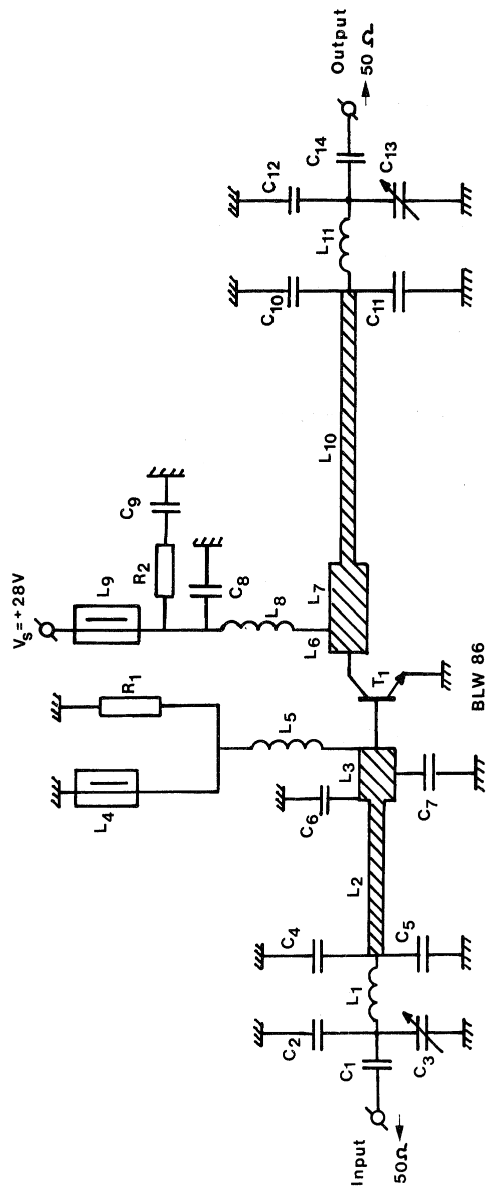

Fig.12 circuit diagram

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PARTS LIST

R_1=12.1Ω metal film. Philips MR 25 cat no.2322 151 71219

R_2=10 Ω metal film. Philips MR 25 cat no.2322 151 71009

C_1=C_8=C_{14}=2k7 chip.Philips NPO size 1210 cat.no 2222 852 13272

C_2= 33pF chip.ATC 100 B-330-J-Px-500

C_3=C_{13}=2-18pF film dielectric trimmer Philips cat no 2222 809 09003

C_4=C_5=120pF chip ATC 100B-121-J-Px-300

C_6=C_7=510pF chip ATC 100B-511-M-Px-100

C_9=100nF metallized film capacitor Philips cat no 2222 352 45104

C_{10}=C_{11}=30pF chip ATC 100B-300-J-Px-500

C_{12}=18pF chip ATC 100B-180-J-Px-500

L_1= 48nH 4 turns enamelled cu-wire Ø 0.8 mm, i.d. 3 mm,
closely wound, length 3.5mm,leads 2x5 mm.

L_2= 60.2Ω stripline w=2 mm, l=27.2 mm

L_3= 30.1Ω stripline w=6 mm, l= 7.9 mm

L_4=L_9 FXC 3B r.f. choke Philips cat no.4312 020 36640

L_5=200nH 14 turns enamelled Cu-wire Ø 0,5mm, i.d. 3 mm,
closely wound, length 9 mm.

L_6=30.1Ω stripline w=6 mm, l= 3 mm.

L_7=30.1Ω stripline w=6 mm, l= 11.8 mm.

L_8=27.9nH 4 turns enamelled Cu-wire Ø 1 mm, i.d. 4 mm.
length 14.3 mm, leads 2x5 mm

L_{10}= 60.2Ω stripline w= 2 mm l= 47 mm

L_{11}=55nH 4 turns enamelled Cu-wire Ø 1 mm, i.d. 4 mm.
length 5.5 mm, leads 2x5 mm

T_1= BLW 86

P.C. board material is epoxy fibre-glass (ε_r=4.5)
thickness 1.6 mm.

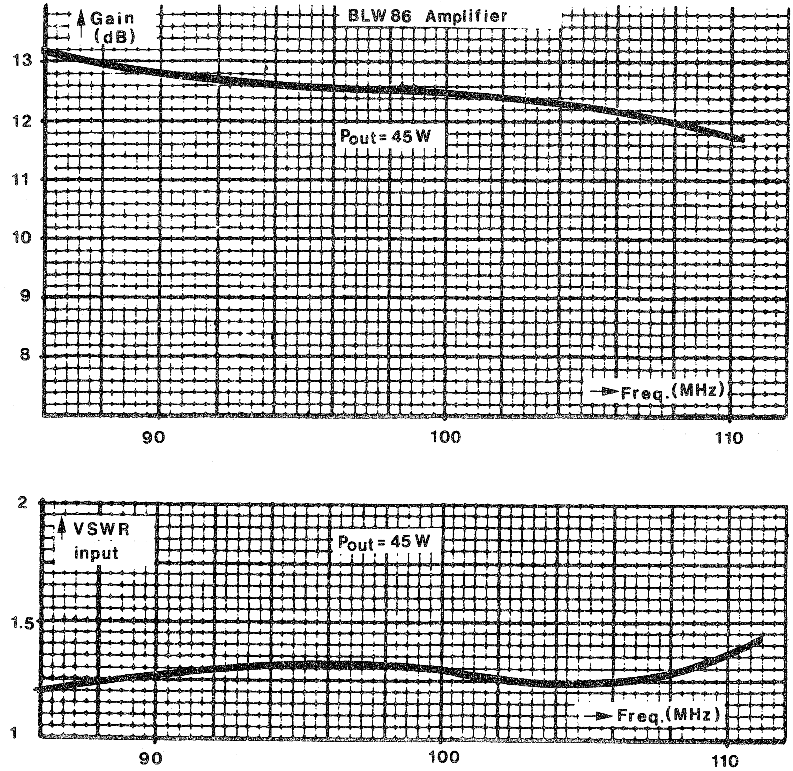

Fig.13 gain and input VSWR versus frequency

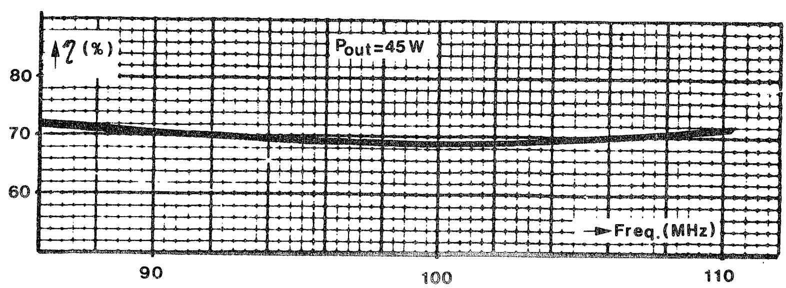

Fig.14 collector efficiency η versus frequency

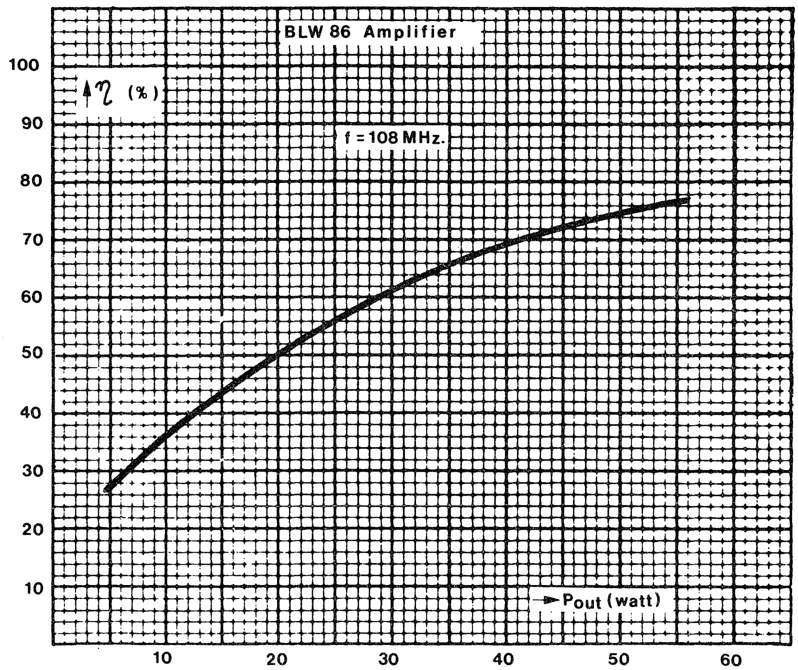

Fig.15 collector efficiency versus output power

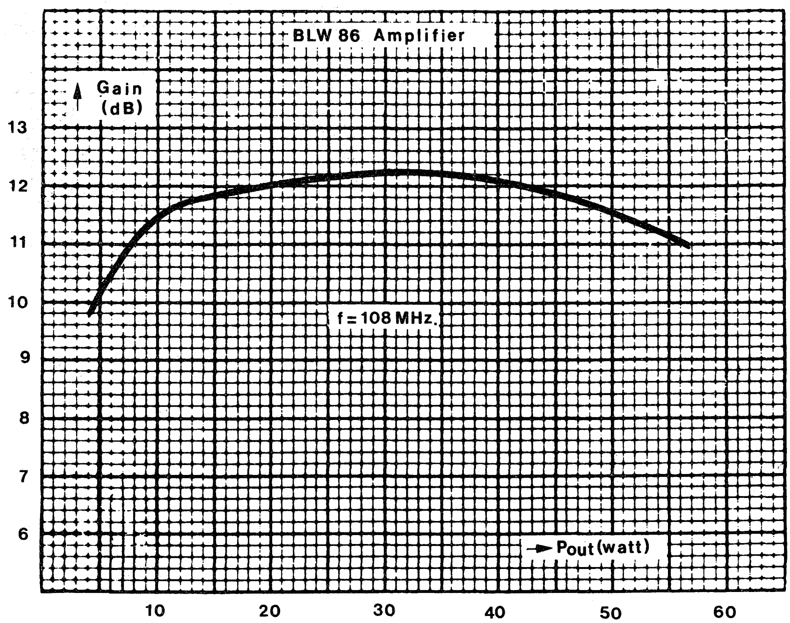

Fig.16 gain versus output power

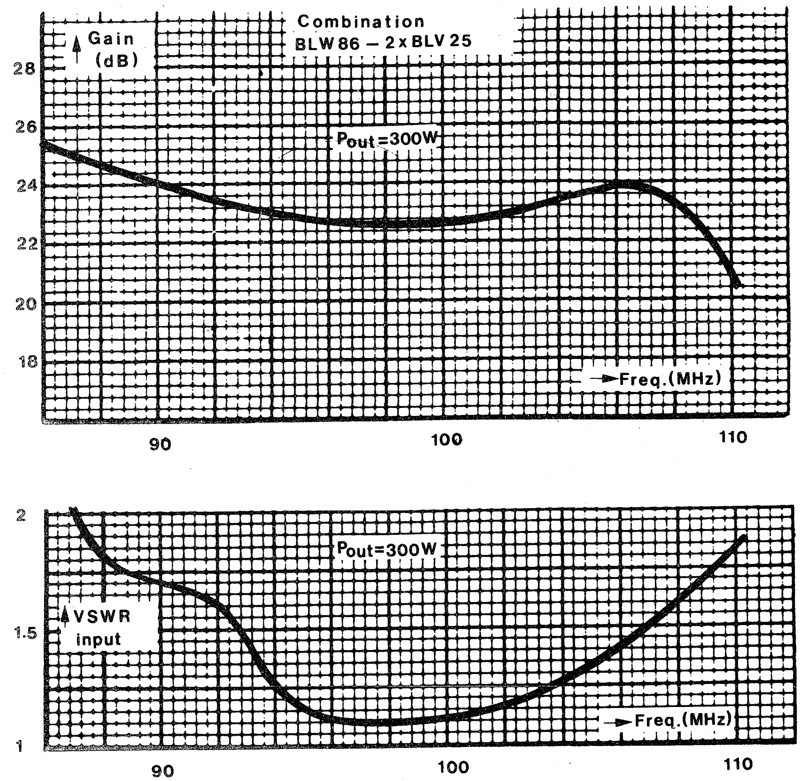

Fig.17 gain and input VSWR versus frequency

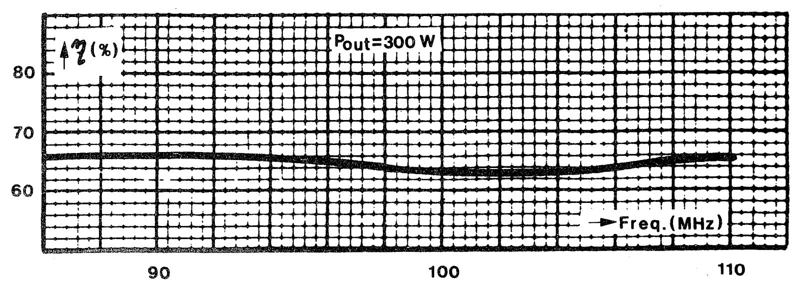

Fig.18 efficiency versus frequency

Electronic components and materials

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N.V. PHILIPS SEMICONDUCTORS APPLICATION LABORATORY
NIJMEGEN – THE NETHERLANDS

REPORT No: NCO 8204	AUTHOR: Mr. H. v. Hees
PROJECT No:--	DATE: 27-12-1982

TITLE

IMPROVEMENT OF STABILITY AND EFFICIENCY OF THE 2 x BLV 25 WIDEBAND AMPLIFIER FOR THE FM BROADCAST BAND (87.5 - 108 MHz)

In report NCO 8202 d.d. 17-05-1982, Mr. v. Hees described a 300 watt wideband amplifier with two transistors BLV 25 for the FM broadcast and (87.5 - 108 MHz). We have built three amplifiers, two on a heat-ink and one with a water-cooling system.
It appeared that with some transistors, it is possible that, at reduced output-power, instability occurs. An effective means to avoid this instability is the application of an inductance L_{cc} between the collectors of the transistors. (See Fig. 1).

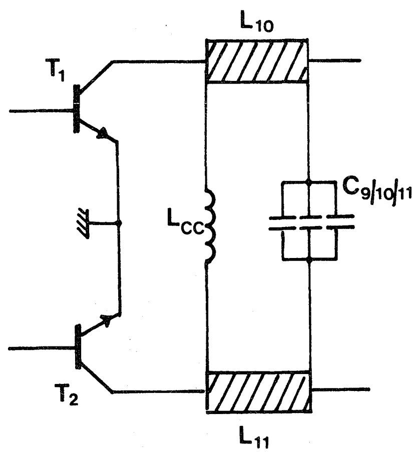


Fig. 1.

Advice Patents Dept. d.d:	AV	GV		B		BL	
Decision MAMO d.d:	AV	GV	EI	B		BL	
DATE:	MAMO:						

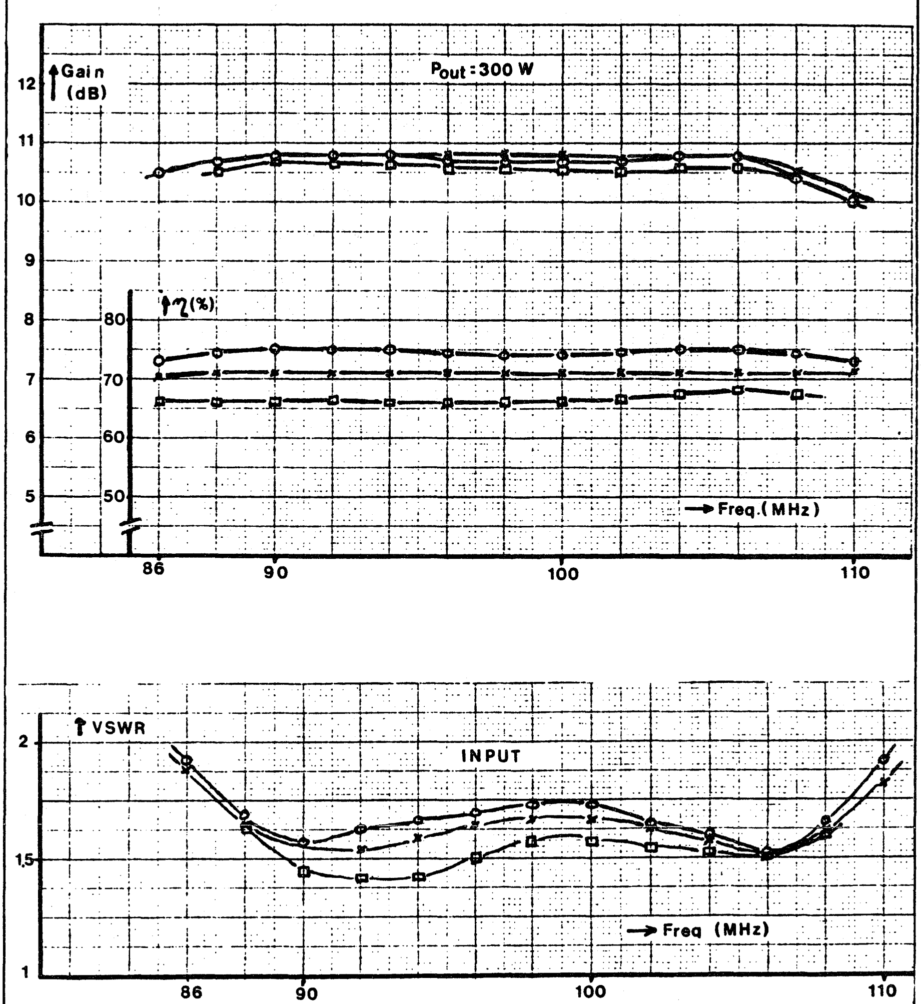
An additional advantage of this inductance is the improvement of the collector efficiency.
In fig. 2 are the results of measurements on a water-cooled amplifier in three conditions, viz. L_{cc} = not present, L_{cc} = 41 nH and L_{cc} = 29 nH.

Conclusions :
With L_{cc} present :
- No instability occurs at any outputpower level at all.
- Collector efficiency improves considerably.
- Input VSWR worsens, however, remains below 1.75.

H. v. Hees

PHILIPS

Electronic components and materials



□-□	No L_{CC}	
x-x	L_{CC}=41nH;	2 turns enamelled Cu-wire Ø 1.7mm, int.diam. D=8mm, length 6mm, leads 2x10mm
o-o	L_{CC}=29nH;	1 turn enamelled Cu-wire, Ø 1.7mm, int.diam. D=10mm, leads 2x12mm.

Fig.2

Semiconductor application laboratory SAL Nijmegen - The Netherlands

Report no.: RNR-1-320-1986-AS / NCO 8601
Author : A.Hilbers
Date : 1986-06-05

RF POWER MOS-TRANSISTORS FOR THE HF AND VHF RANGE

1. INTRODUCTION

In the past 2 decades PHILIPS have introduced more than 100 types of bipolar transmitting transistors.
For most applications these devices show a satisfactory performance.
There are however areas where a better performance concerning noise is required, e.g. in duplex equipment or in cases where many transceivers in the same area are in operation.
For this purpose PHILIPS have started now the introduction of a series of RF power MOSFETS. In the beginning this is concentrated in the HF and VHF range, i.e. frequencies below 250MHz. The table below gives a survey of the present situation.

Type	f(MHz)	V_d(V)	P_o(W)	d_3(dB)	Encaps.	Status
BLF 242	175	28	5	-	SOT 123	Dev.
BLF 244	175	28	15	-	SOT 123	Prod.
BLF 245	175	28	30	-	SOT 123	Prod.
BLF 145	28	28	8	-40	SOT 123	Dev.
BLF 146	28	28	80	-30	SOT 121	Dev.
BLF 147	28	28	150	-30	SOT 121	Dev.
BLF 177	28	50	150	-30	SOT 121	Dev.

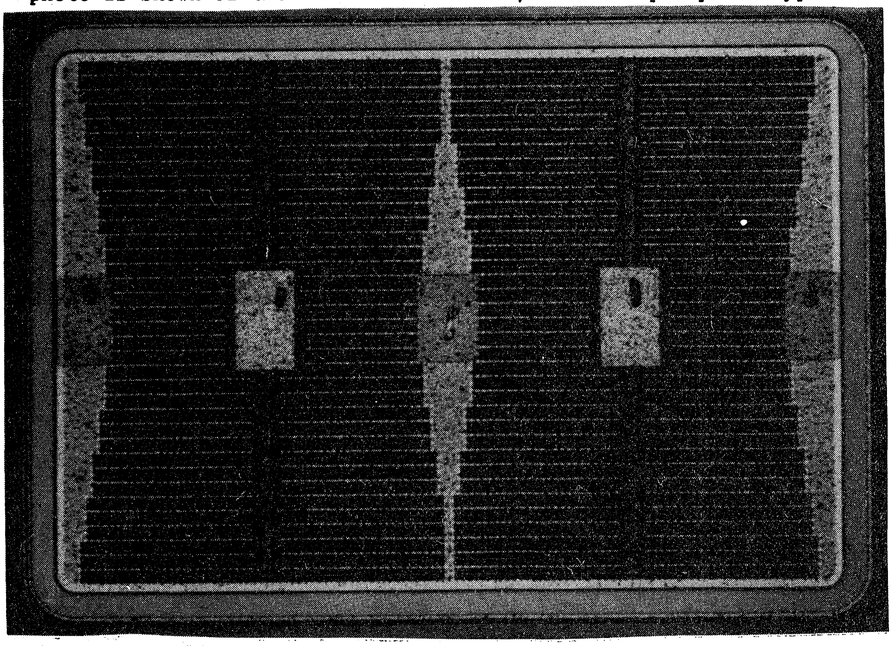
In the future this range will be extended towards higher frequencies, lower and higher output powers and supply voltages and different encapsulations.
Coming back to the noise improvement mentioned above this was appr. 7dB measured at 75MHz in similar wideband amplifiers. Other advantages are:
- Higher power gain than comparable bipolar types (4-5dB).
  This is mainly due to the high transconductance and low feedback capacitance.

- Control of the output power down to almost zero is simply possible by reducing the gate-source DC voltage.
- Thermal stability of a MOSFET is principally better than that of a bipolar transistor thanks to the negative temperature coefficient of the drain current at higher levels.
  This is also the reason for the good current distribution over the whole active area of the device.
- As a consequence the load mismatch capability of a MOSFET is better than that of a comparable bipolar type.

Ofcourse there are also disadvantages like:
- Sensitivity of the gate for charges
- Some form of gate bias network is always required at higher frequencies.
- output power slump,i.e. drop versus temperature is more for a MOSFET than for a bipolar type. This is not a matter of reduced saturation power but a reduction of power gain caused by a decreasing transconductance.

2. CONSTRUCTION OF RF POWER MOSFETS

Encapsulation and bonding are the same as for a bipolar type.Below a photo is shown of the X-tal of a BLF 244, a 15W output power type.



It is made on epitaxial material with a low-ohmic N+ substrate and an N-layer with well defined thickness and resistivity. As one can see the structure is of the interdigitated type with 3 bonding pads for the source and 2 for the gate. The X-tal size is appr.1.4x1.0mm^2.

The drawing below shows a detail of the cross-section of this structure. It is essentially a vertical D-MOS device with a metallized poly-silicon gate. The operation of this type of structure will not be discussed here because it can be found in several books and articles. The electron current starts from the N+ source region, continues its way through the upper part of the P-region (the channel) and after having reached the N-region (the drain) it bends downwards in the direction of the N+ substrate.

RF-VDMOST

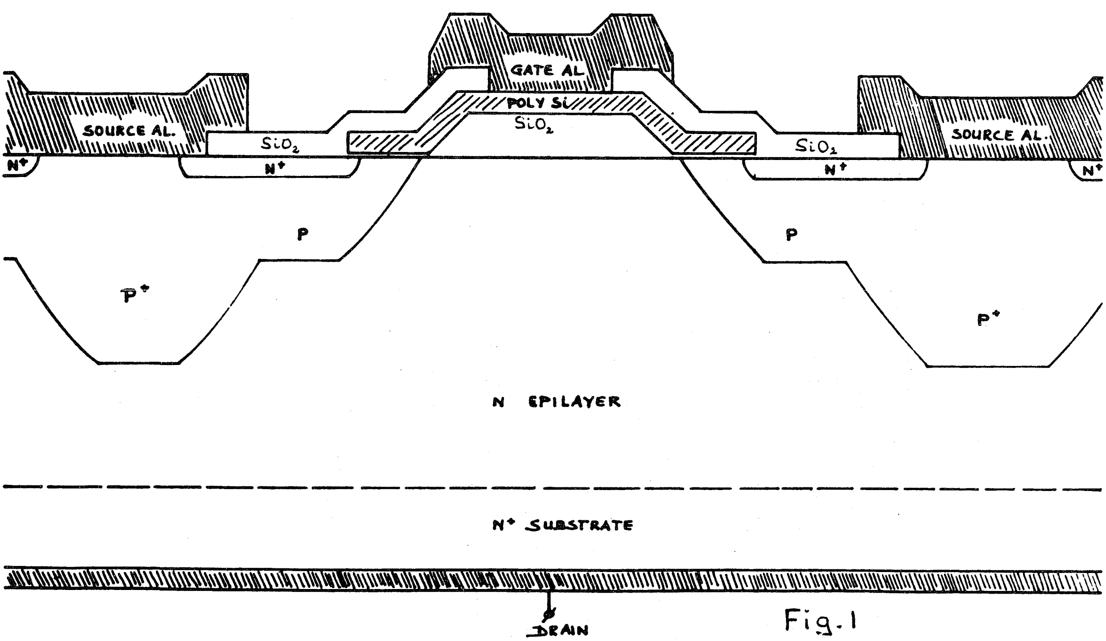


Fig.1

The main advantages of this structure are:

- its large amount of channel width (80-90 mm) per square mm of active area. The channel width is the dimension perpendicular to the drawing. It accounts for the amount of drain current and output power the transistor can deliver. In this way a high transconductance and a low $R_{DS(on)}$ can be realized. The old story that a MOSFET delivers less output power per unit of area compared to a bipolar type is no longer true for this structure. An additional feature is its excellent linearity (low IM distortion) because the high transconductance is maintained up to high drain current levels.
- the feedback capacitance has been reduced by increasing the oxyde thickness above the drain area. Together with the high transconductance this leads to increased stable gain at high frequencies.

3. DC AND RF PROPERTIES

3.1. Ratings

The channel region which is electrically connected to the source forms a diode with the drain region. The avalanche breakdown of this diode is responsible for the V_{DS}-rating.

The V_{GS} rating is determined by the thickness of the gate oxyde.

This rating must never be exceeded because this causes permanent damage to the transistor. Reasonable precautions in handling should be observed to protect the device from electrostatic charge.

The drain current ratings are based on the maximum current the device can deliver without too much reduction of the transconductance.

As with bipolar transistors the thermal resistance increases as a function of power dissipation and mounting base temperature. This is because of the fact that the same materials i.e. silicon and beryllia have been used.
However because of the very good current distribution over the crystal there is no significant difference in R_{th} between DC and RF operation; there is also hardly any dependence on drain voltage for the same dissipation.

3.2. DC characteristics

Both the drain and gate leakage currents of MOSFETS are extremely small; they are typically in the nano-ampere range.

Fig.2 shows the I_D versus V_{GS} characteristics of a BLF 244 at junction temperatures of 25°C and 125°C.

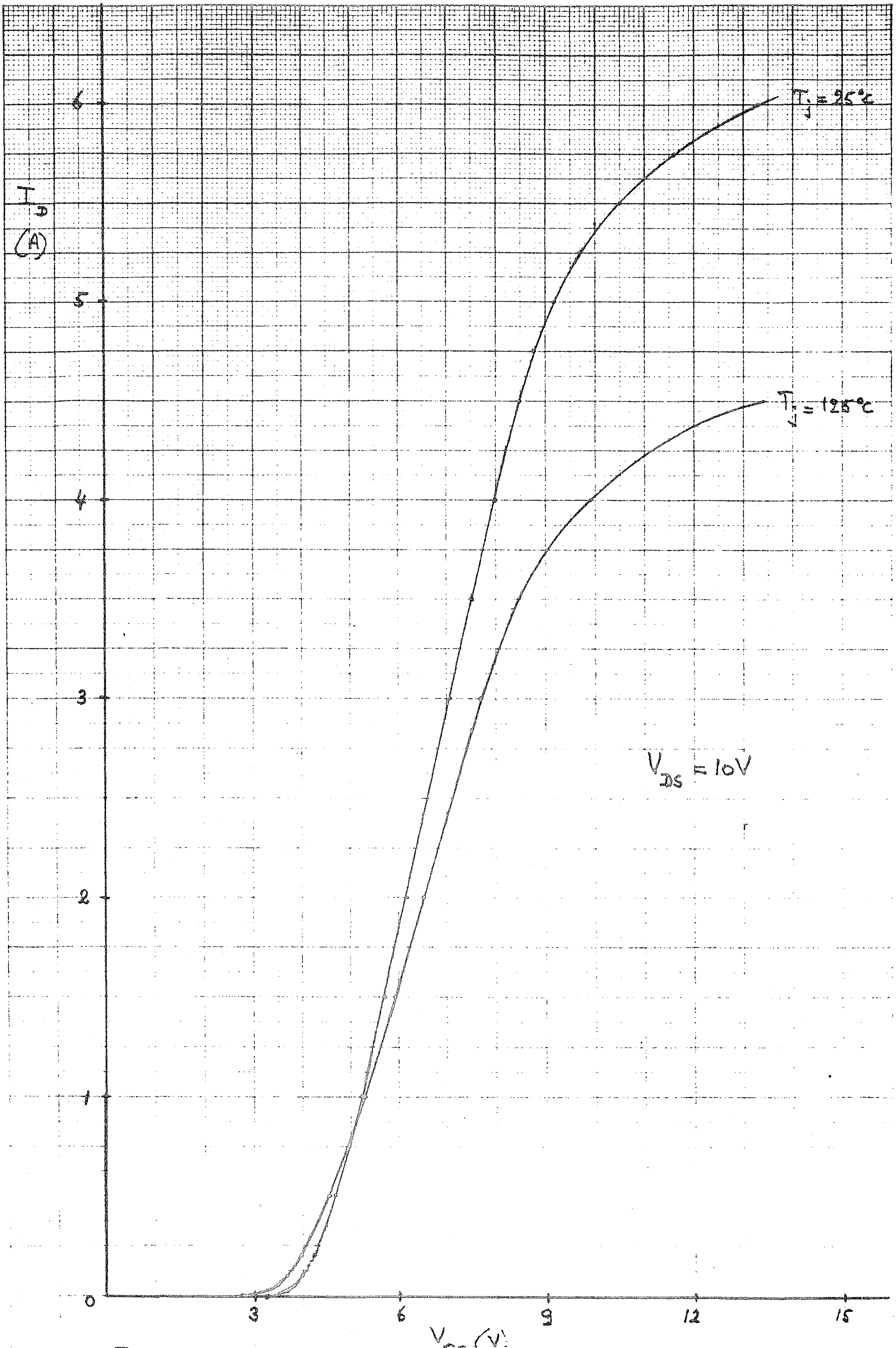


Fig. 2

BLF 244

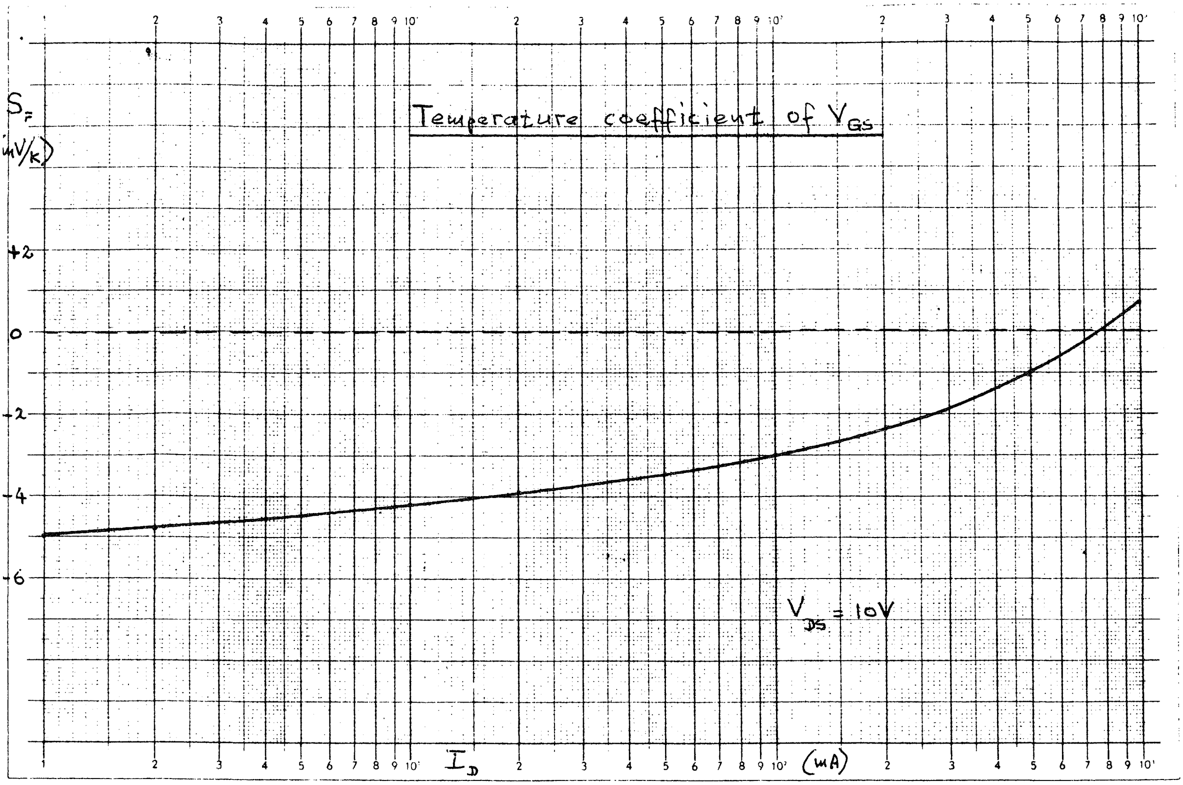
As long as V_{GS} is below appr. +3V the drain current is almost zero. Above that voltage conductance starts and the drain current is then appr. proportional to the square of V_{GS}. This continues up to an I_D of appr.1A. Above that current the relationship between I_D and V_{GS} is nearly linear up to an I_D of appr.4A. For higher drain currents the transconductance reduces gradually to much lower values than in the linear part.

When the junction temperature is increased the whole curve turns clockwise around a point corresponding with an I_D of 0.7-0.8A. This happens to be the normal operating current of the device. If a device is used in class-A at this current there is no need for any temperature compensation of the bias unit. In class-AB where I_{DQ} is much lower, temperature compensation is certainly needed. As the required amount can not be read very well from the graphs of fig.2 another curve is presented (see fig.3) giving the temperature coefficient of V_{GS} needed to stabilize I_D.

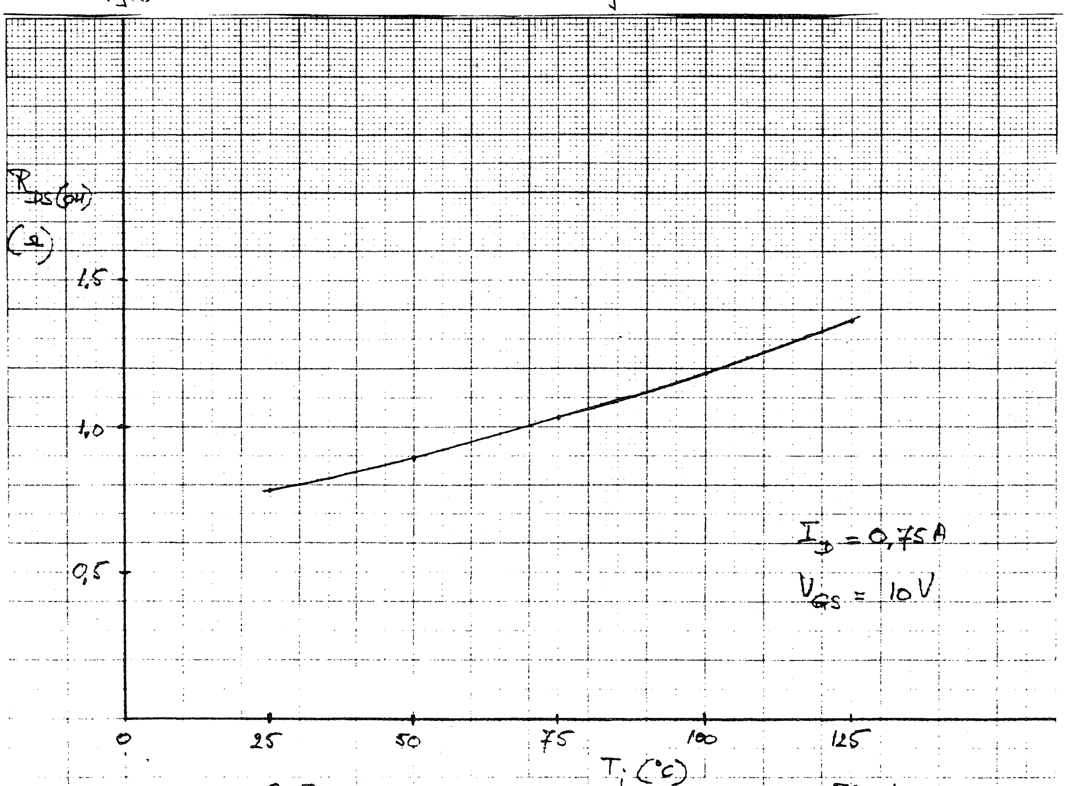
The inversion of the temperature coefficient at higher values of I_D assures that thermal runaway will not occur for a well heatsinked device.

The transconductance is inversely proportional to the absolute temperature raised to a power between 0.9 and 1.0.
This will cause a power gain reduction of amplifiers at higher temperatures.

Fig.4 shows the $R_{DS(on)}$ as a function of T_j. It is appr. proportional to the square of the absolute temperature.



meetpapier - wormerveer Fig.3 BLF 244 No. 17 H $T_j = 25-125°C$ X-as log. verdeeld 1-10' Eenheid 90 mm Y-as verdeeld in mm



BLF244 Fig. 4

The V_{GS} at which conduction starts, the so-called threshold voltage, spreads from 2.0 to 4.5V.
In balanced amplifiers it is an advantage if one common bias unit can be used. For this purpose PHILIPS deliver matched pairs.
This matching is done on basis of a maximum V_{GS} difference of 100mV.
However it must be mentioned that V_{GS}-groups can not be ordered individually.

3.3. RF characteristics

In our publications we give 3 capacitances, viz.:
C_{iss}, C_{oss} and C_{rss}. For device modelling purposes it is useful to know that:

$C_{GS}=C_{iss}-C_{rss}$

$C_{DS}=C_{oss}-C_{rss}$

$C_{GD}=C_{rss}$

Fig.5 shows the capacitances as a function of V_{DS}.
C_{oss} is mainly formed by the reverse biased drainsource diode.
It shows a large variation.
This is also the case with the feedback capacitance C_{rss}.
The input capacitance (C_{iss}) however shows a rather constant behaviour.

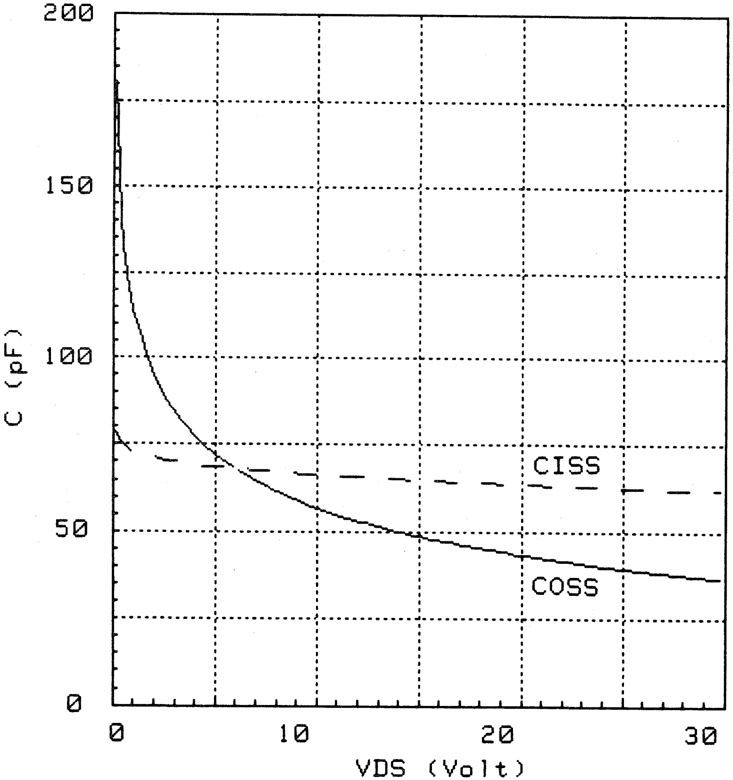
Although it is somewhat unusual it is also for a MOSFET possible to define an f_T. This is equal to:

$$G_{fs}/(2\pi C_{iss})$$

For a BLF 244: G_{fs} = 0.87S (at I_D=0.75A)
C_{iss} = 59pF (at V_{DS}=28V)

So f_T = 2.35 GHz.

BLF 244; $V_{GS} = 0$; $f = 1$ MHz



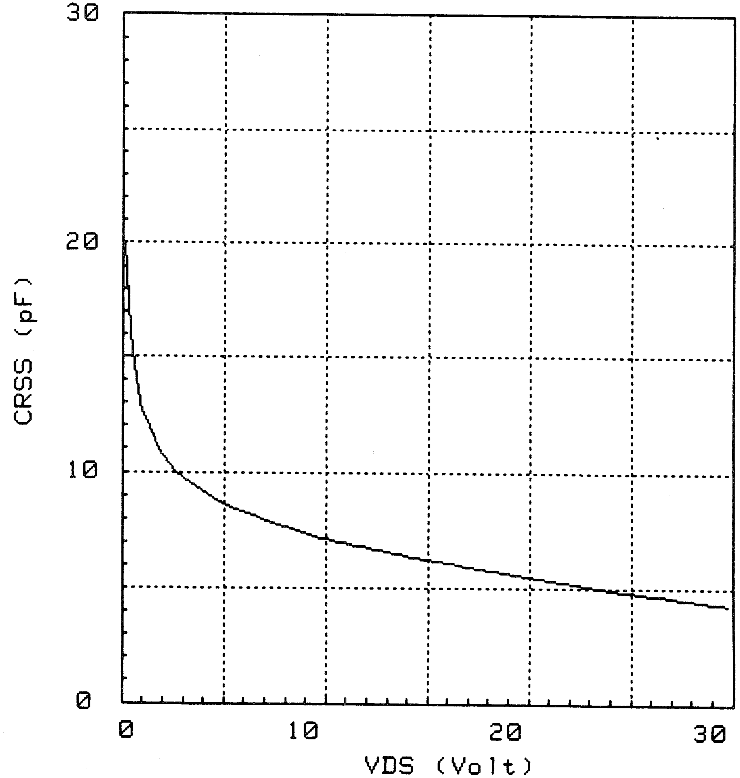


Fig. 5

6a

S-parameters have been measured on some samples of BLF 244 in a different encapsulation (SOT 122 instead of SOT 123). The DC adjustment was: V_{DS}=28V; I_D= 0.5A.

The results are shown in Table I.
As MOSFETS are in general not unconditionally stable it is better to judge their performance on basis of the MSG= maximum stable gain.
It can be concluded that these devices have useful power gains up to at least 500MHz.
It must be remarked that the actual BLF 244 (in SOT 123) will have a somewhat lower power gain.

The noise figures mentioned in our publications are measured in the published test circuits under power match conditions. For a BLF 244 this is done at f=175MHz; V_{DS}=28V and I_D=0.5A.

The result is:
F= 4.3-4.6dB.
In this case a gatesource damping resistor of 23 Ohms was used.
Experiments with BLF 244 and BLF 245 in circuits with high R_{GS} values and matched at the input for minimum noise figure have shown values between 1.5 and 2.6dB.
In those cases the power loss of the input circuit has been estimated at 0.5dB so that the actual device noise figure must have been between 1 and 2dB. However these values can never be obtained in practical amplifiers.

4. POWER AMPLIFIERS

4.1. Gate biasing

Power gain and drain efficiency are measured in testcircuits where the drain quiescent current (I_{DQ}) is adjusted for each transistor individually by means of V_{GS}. As long as there are no requirements on linearity like IM distortion there is nothing against operation from a fixed V_{GS} below the threshold voltage, e.g. 2.0V. In fact this is class-C operation and the result is a somewhat smaller powergain but improved drain efficiency.

7a

S-PARAMETERS $V_{DS}=28V;\ I_D=0.5A$

f	S11		S21		S12		S22		GUM	MSG
(MHz)	M(dB)	Ang.	M(dB)	Ang.	M(dB)	Ang.	M(dB)	Ang.	(dB)	(dB)
10	-1.0	- 69	32.8	137	-31.8	50	-1.8	- 61	44.4	32.3
20	-1.9	-106	29.0	115	-29.3	31	-3.4	- 96	36.1	29.1
50	-2.3	-144	22.0	94	-28.0	11	-4.3	-134	27.9	25.0
100	-2.2	-161	16.2	79	-28.5	0	-4.0	-149	22.4	22.4
200	-1.7	-170	9.5	57	-31.6	0	-2.9	-156	17.4	20.5
300	-1.2	-175	4.8	41	-34.5	28	-2.0	-161	15.3	19.6
400	-0.9	180	1.1	29	-32.3	62	-1.3	-165	14.0	16.7
500	-0.8	175	-2.0	21	-28.7	76	-0.9	-169	13.1	13.3

GUM = Maximum unilateralized gain

MSG = Maximum stable gain

BLF 244 in SOT 122 encapsulation

TABLE I

Another alternative is the use of a Schottky-barrier diode in parallel with R_{GS}. In that case no external V_{GS} is required because it is generated by the diode. This diode must be connected with its anode to ground. Suitable types are: BAT81 and BAT82.
R_{GS} must have a high value, e.g. 100K Ohm. If RF damping between gate and source is required the circuit configuration of fig.6 can be chosen.

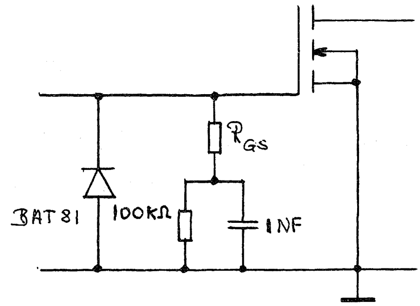


Fig.6

This solution leads also to a somewhat lower powergain and higher drain efficiency.

4.2. Powergain, input and load impedance

For the main application area these quantities will be taken up in the publication. At lower and medium frequencies it is quite well possible to calculate these quantities with reasonable accuracy. This can be done by means of an equivalent circuit diagram. As an example we take the BLF177, a 150W, 50V transistor for the HF-SSB range of 1.6 to 28MHz. In fig.7 an equivalent circuit diagram of an RF power MOS transistor is given.
We will start with the optimum load impedance. For the use at lower frequencies we may simplify the diagram of Fig.7.

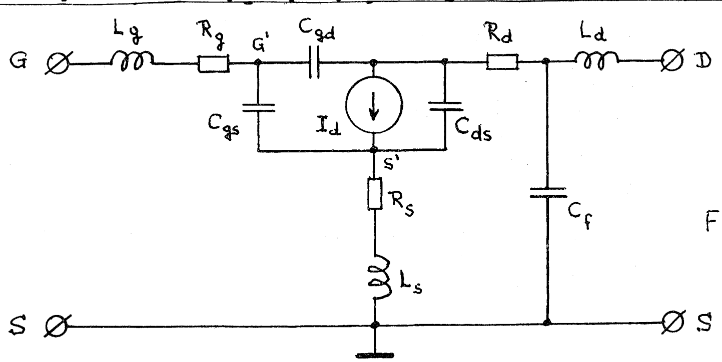


Fig.7

See fig.8. R_L can be determined with:

$$R_L = V_d^2/(2P_o),$$

in which V_d is the amplitude of the drain AC voltage and P_o the corresponding outputpower.

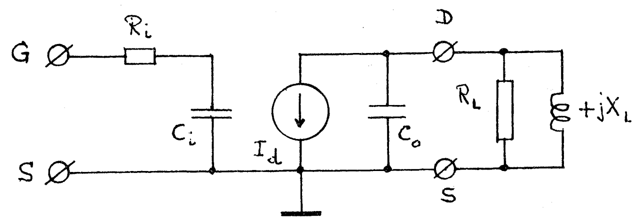


Fig. 8

As linear operation is required V_d must be chosen lower than the 50V supply voltage. Taking 44V (see section 6.1) we get:

$R_L = 44^2/(2.150) = 6.5$ Ohm

Because of the large drain voltage swing C_o is appr.15% higher than the published C_{oss}, so:

$C_o = 1.15 \times 190 = 218$pF

This capacitance must be tuned out with an inductive reactance (parallel or series equivalent).
The current generator I_D equals:

$I_D = G_{FE} \times V_{GS}$

in which G_{FE} is the effective transconductance.In class-AB with a relatively small quiescent drain current G_{FE} is 50% of the published G_{FS}. In addition to this comes a reduction of appr. 25% due to the junction temperature rise caused by normal power operation and being appr.100^{o}C. With a published G_{FS} of 6.2S we get:

$G_{FE} = 0.5 \times 0.75 \times 6.2 = 2.3$S.

The voltage gain is:

$A = G_{FE} \times R_L = 2.3 \times 6.5 = 15.$

The appr. device input impedance can be determined as follows:

$C_i = C_{gs} + C_{gd}(A+1).$

The second term is caused by the well known Miller-effect.
For the BLF 177:

$C_{iss} = 470pF$

$C_{rss} = 19pF$

So $C_{gs} = 470-19=451pF$.

Because of the large drain voltage swing it is assumed that $C_{gd}(=C_{rss})$ is increased by 15%, so $C_{gd}= 1.15x19=22pF$.

The total input capacitance becomes then:

$C_i=451 + 22 (15+1) = 803pF$.

The resistance R_i is for the larger part determined by the source inductance L_S. It can be proven that:

$R_i= w_{T(eff)} x L_S= G_{FE} x L_S/C_i$

The calculated L_S of the BLF 177 is 0.72nH, giving:

$R_i = 2.3 x 0.72E-9/803E-12=2.1$ Ohm

The equivalent parallel resistance of the input at 28MHz is then 26 Ohms.

For several reasons on which we come back later the input will be shunted with a resistor of 6.25 Ohms.
This resistor dominates the power gain which is:

$G_p = 10 x LOG (G_{FE}^2 x R_L x R_{GS})= 10xLOG (2.3^2x6.5x6.25)=23.3dB.$

The actually measured power gain is 22.6dB.
As one can see the difference is small. It can be explained by losses in the matching networks, in R_d (0.2-0.3 Ohm) and in the above mentioned input parallel resistance of 26 Ohms.

4.3. Stability

As mentioned before MOSFETS are in general not unconditionally stable. The losses in the gate-source circuit are exceptionally small so that a detuning of the load in the inductive direction easily leads to oscillation except when sufficient damping is present between gate and source.
It can be proven that the circuit is stable if:

$wC_{gd}G_{FE}R_LR_{gs}<2$. (See e.g. Ref.1).

A possible stabilizing effect from L_S has been neglected here.

The maximum stable gain (MSG) is then:

$MSG=10xLOG (G_{FE}/(w x C_{gd}))$ dB

Because of the above our test circuits have been restricted in the tuning range of the output network by applying only one variable capacitor. In other circuits for linear applications the output network is tuned by means of a dummy load (for calculation of the components see e.g. the previous section).
Another measure taken in most cases is of course a gate-source resistor of sufficiently low value.

4.4. Matching networks

Input and output impedance matching networks for MOSFETS are not principally different from those for bipolar types. Therefore they will not be described here. One is referred to the numerous reports written on this subject by PHILIPS and others.

4.5. Power saturation

The maximum drain current of our MOSFETS is rather high compared to what is required for the application. A factor 2 is quite normal. Thanks to this fact their linearity is excellent. There is however also a danger and that is that most of the devices are able to deliver twice the published power (and sometimes more). We do not advise to exploit this fact because long operating life can only be guaranteed if the devices are used at or below their published power levels. Exceptions are probably pulse operation, amplitude modulated amplifiers and the like.

4.6. Load mismatch capability

This is published for all our types. It is equal or better than for comparable bipolar types because of improved current distribution. The margin between the publication value and the actual point of destruction is larger for low power types in comparison with high power types. The increase of drain current during mismatch is relatively small due to the negative temperature coefficient of I_D at higher values.

4.7. Power slump

By this we mean the reduction of output power with increasing heatsink temperature. When the latter is increased from 25^o to 70^oC the power slump is appr.0.5dB for the BLF 244 and BLF 245. It is a little bit more than for comparable bipolar types.
As mentioned before this power slump is mainly caused by reduction of transconductance versus junction temperature. Efficiency and saturation power are hardly influenced so that correction is rather easy.

4.8. Control of output power

The output power of a MOSFET can easily be controlled down to almost zero by reduction of the gate-source bias voltage. In normal operation the BLF 244 and BLF 245 need a bias voltage between +3 and +4V.
For power control down to zero this voltage must be reduced to -5 to -6V at 175Mhz.
The advantage of this way of power control is that hardly any control power is needed. In this way amplitude modulation becomes a rather simple matter.

5. WIDEBAND APPLICATIONS

5.1. Restrictions at the output side

The output bandwidth of a transistor is appr. determined by:

$$B_o = 1/(2\pi R_L C_o)$$

in which R_L is the load resistance (parallel component) and C_o the effective output capacitance.
Due to the drain voltage swing the latter is appr.15% more than the published C_{oss}. Let's give an example.
For the BLF 245:

$$R_L = V_d^2 /(2P_o) = 28^2/(2x30) = 13.1 \text{ Ohm}$$

$$C_o = 1.15\ C_{oss} = 1.15 \times 76 = 87\text{pF}$$

So: $B_o = 1/(2\pi\ 13.1x87E-12) = 140\text{MHz}$

This is the maximum bandwidth that can be obtained without sacrificing much output power and drain efficiency. Larger bandwidths can be obtained at the cost of more drop in power and efficiency.
The figure for B_o given above is somewhat worse compared to similar bipolar types, however it is still more than adequate for most HF and VHF applications.

5.2. Restrictions at the input side

The bandwidth at the input side of the transistor is:

$$B_i = K/(2\pi R_{GS} C_i)$$

in which K is a factor depending on the system of input matching.
For simple systems K is 1 but for more sophisticated ones it can be as high as 1.6 as we will see in the next section.
R_{GS} is the externally connected gate-source damping resistor.

So the input bandwidth is inversely proportional to the value of this resistor.
As we have seen earlier the power gain (as a ratio) is proportional to the value of R_{GS}. This means that the power gain bandwidth product is a constant as long as it is not restricted by the output.

5.3. Example of a practical wideband amplifier

On page 13a the schematic diagram and parts list are given of a wideband power amplifier for military communication purposes. It has been designed around the BLF 245 and it delivers 25-30W at a supply voltage of 28V. The frequency range is from 25 to 110MHz.

As we have seen in section 5.1. the optimum load resistance for this type is 13.1 Ohm. For the ease of transformation we have chosen 12.5 Ohm. This is obtained by a 1:4 impedance transformer of the transmission line type with a ferrite core. The usual LF and HF compensations have been applied. The purpose of L6 together with a part of C11/C12 is to compensate the effect of the transistors output capacitance.

At the input side we have to cope with an effective input capacitance of 220pF (measured). The input matching system chosen is depicted in fig.9.
C_i represents the input capacitance of the BLF 245 across which we have to develop a constant voltage versus frequency from 25 up to 110MHz. Provided C_i is an ideal capacitor, i.e. there

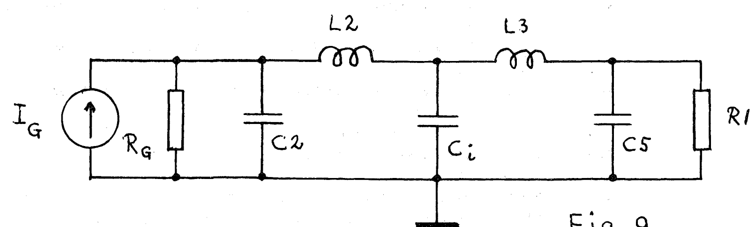


Fig. 9

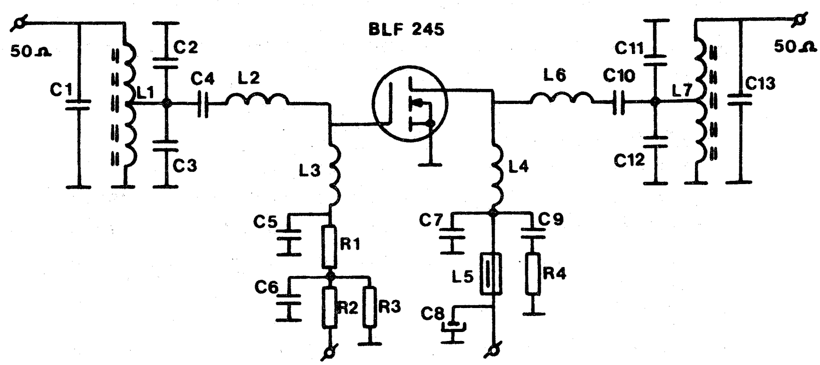
are no series or parallel losses, the optimum dimensioning of this network follows from the following formulae:

$R_G = R_1 = 1.6/(w_c . C_i)$

$C_2 = C_5 = 0.386\ C_i$

$L_2 = L_3 = 0.997\ R_1/w_c$

in which w_c is the maximum angular frequency.



WIDEBAND POWER AMPLIFIER WITH BLF 245 (f=25-110 MHz)

C1 = 8.2pF multilayer ceramic chip capacitor *
C2-C5 = 100pF multilayer ceramic chip capacitor *
C3 = 62 pF multilayer ceramic chip capacitor *
C4=C10= 10nF multilayer ceramic chip capacitor (cat.nr.2222 852 47103)
C6=C7= 100nF multilayer ceramic chip capacitor (cat.nr.2222 852 47104)
C8= 2.2uF electrolytic capacitor
C9= 3x100nF multilayer ceramic chip capacitor(cat.nr.2222 852 47104)
C11= 82pF multilayer ceramic chip capacitor *
C12= 43pF multilayer ceramic chip capacitor *
C13= 12pF multilayer ceramic chip capacitor *

L1= 2x Ferroxcube toroids, grade 4C6 (6x4x2mm)(cat.nr.4322 020 97160) with 6 turns of 2x0.25mm twisted enamelled Cu-wire (See fig. 1)
L2= 17.6nH, 2 turns enamelled Cu-wire (0.6mm) int.dia.:3mm ,length 2.5mm, leads 2x5mm
L3= 28.8nH, 3 turns enamelled Cu-wire (0.6mm) int.dia.:3mm ,length 3.2mm, leads 2x5mm
L4= 455nH, 12 turns enamelled Cu-wire (1mm) int.dia.:7mm ,length 16.5mm, leads 2x5mm
L5= Ferroxcube h.f.choke,grade 3B (cat.nr.4312 020 36642)
L6= 10nH,1 turn enamelled Cu-wire (1mm) int.dia.:3mm leads 2x3mm
L7= Ferroxcube toroid,grade 4C6 (23x14x7mm) (cat.nr.4322 020 97190) with 5 turns of 2x0.7mm twisted enamelled Cu-wire (See fig. 1)

R1= 12.4 Ohm,parallel connection of 5 metal film resistors 61.9 Ohm (cat.nr.2322 151 76199)
R2= 1K Ohm, metal film resistor (cat.nr. 2322 151 71002)
R3= 1M Ohm, metal film resistor (cat.nr. 2322 151 71005)
R4= 10 Ohm, metal film resistor (cat.nr. 2322 153 51009)

PC-board: double Cu-clad, 1.6mm epoxy fibre-glass (ε_r=4.5)

* American Technical Ceramics type 100B or capacitor of same quality.

This has been found with a computer optimization program which also indicates that the maximum voltage variation across C_i is +/- 0.36dB and the maximum input VSWR seen by the generator is 1.36.
In our case we find that:
$R_G=R_1=10.5$ Ohm
$C_2=C_5=85pF$
$L_2=L_3=15.1nH$
For the ease of transformation R_G and R_1 have been chosen 12.5 Ohm.
Further it is so that the resistive component of C_i is substantial certainly in the upper part of the frequency band. Therefore some of the component values had to be changed to get a maximally flat response. The result of this can be seen in the parts list.
The remaining part of the transformation from 50 Ohm to 12.5 Ohm has been accomplished with a transformer very similar to the one used at the output, however it is wound on a smaller core and connected in the inverse way.
The most important results obtained with this amplifier are;

P_o= 27.5W G_p = 17.7 +/- 0.5dB
V_d= 28V η_D= 53-67%
f = 25-110MHz

The maximum input VSWR at the high end of the frequency band is 1.67. This is higher than predicted due to:
1. the choice of R_G and R_1
2. the damping of C_i

Looking at the power gain we have the following situation.
G_{FS} is typ. 1.7S for this type at T_j= 25°C
In class-B at T_j= 125°C it reduces to appr.0.64S (see section 4.2).
For the transistor the generator and load resistance are both 12.5 Ohm, so:

G_p=10x LOG $(0.64^2$x 12.5 x 12.5)= 18.1dB.

The measured power gain is only slightly lower due to losses in the matching networks and in the transistor ($R_{DS(on)}$) which were not taken into account.

6. LINEAR APPLICATIONS

6.1. SSB output stages

Most SSB transmitters operate in the frequency range from 1.5 up to 30MHz. Narrow band tests are carried out at f= 28MHz. Apart from power gain and drain efficiency the 2-tone intermodulation test is very important. This is done with 2 signals of equal amplitude separated in frequency by 1KHz. For output stages the third and fifth order products must be below -30dB when compared with either of the 2 tones.

A good example is the BLF 177 specified for an output power of 150W PEP at a supply voltage of 50V and a frequency of 28MHz. IM distortion is better than -30dB, power gain 20dB min. and 2-tone drain efficiency 35% min.

To obtain this performance in a practical amplifier a number of points is important:

1. the choice of the correct load impedance.
   The resistive part (parallel component) must be:
   $R_L = V_d^2 / (2P_o)$,
   in which V_d must be chosen 85-88% of the supply voltage.
   The reactive part has to tune out the effective output capacitance of the device which is appr.15% higher than the published C_{oss} (see also section 4.2.).
2. the drain quiescent current, i.e. the drain current at zero drive power. For a MOSFET a higher value is required than for a comparable bipolar type. A 50V devic e needs 3-3.5mA per Watt output power and a 28V device 6.5-7mA.
3. the gate-source damping resistor R_{GS}.
   Lower values give better IM distortion at the cost of power gain. A good value for the BLF 177 is 6.25 Ohm. This value also guarantees good stability and an input bandwidth of at least 30MHz.
4. Something must be done to reduce the amplitude of the second harmonic of the drain voltage, otherwise the amplifier will show saturation phenomena before the desired output power has been reached. In a 28MHz narrow band amplifier a drain-source capacitor of 40pF is adequate and in a wideband push-pull amplifier this function can be taken over by a mid-tapped drain choke of which the 2 halves are coupled inductively. This coupling must be as tight as possible.

The influence of heatsink temperature on the performance of linear amplifiers must still further be investigated. The first impression is that IM distortion and drain efficiency are hardly affected and the power gain drop is of the order of 1dB when the heatsink temperature rises from 25^{o} to 70^{o}C.

6.2. SSB driver stages

Driver stages in SSB transmitters must have a lower IM distortion than the final amplifier. A generally accepted figure is -40dB. This can only be achieved with class-A operation. In the near future the type BLF 145 will be introduced for this type of service. Although the exact specification has not yet been made the first measurements give the following indications:

DC operating point: $V_D = 28V$

$I_D = 1.3A$

At f = 28MHz:

$P_o = 8W$ PEP for $d_3 < -40dB$

$G_p > 24dB$

Also here some points must be observed:

1. the optimum load resistance (parallel component) must be slightly less than the ratio of DC drain voltage and current, so: $R_L < 28/1.3 = 21.5$ Ohm.

   We have chosen 20 Ohm.

   For the reactive component the same holds as for output stages (see previous section). This means for this type that an output capacitance (parallel component) of appr. 84pF must be tuned out.
2. The gate-source resistor R_{GS} must be lower than for class-AB operation. This is because of the higher effective transconductance and load resistance leading to a higher voltage gain and more influence of the Miller-effect, i.e. a higher effective input capacitance. For this type we recommend 27 Ohm.

   In wideband amplifiers this resistance can (partly) be replaced by a gate-drain resistor of suitable value.

Concerning the influence of heatsink temperature the first indications are that the output power for a given IM distortion drops by appr.10% when the heatsink temperature increases from 25^{o} to 70^{o}C. At the same time the power gain decreases by appr.1dB.

6.3. TV transmitter service

In TV transmitters vision and sound are separately amplified. The linearity requirement for the vision amplifier is that the gain compression must be less than 1dB. Our BLF 245 has been tested on this point at 175MHz, the lower end of TV band III. It appeared that the gain compression up to the full published output power of 30W was only a few tenth of a dB provided the drain quiescent current was increased from 50 to 200mA. A test at 225MHz has not yet been done but the only difference expected is a somewhatlower power gain. In general MOSFETS follow the 6dB per octave line like bipolar types so the gain reduction from 175 to 225MHz will be appr. 2.2dB.
Drain efficiency and gain compression will hardly change.

For the load impedance of this type of amplifiers the same holds as stated in section 6.1 for class-AB amplifiers in the HF range.

6.4. TV transposer service

In TV transposers vision and sound are amplified together. Therefore the linearity requirements are more severe. The most common test is with 3 tones having levels of -8, -16 and -7dB with respect to a 0dB reference level called the peak sync power. The frequency separation of the first and the last tone is 5.5MHz and the second tone can be anywhere in between.
If the frequencies of the 3 tones are p,q and r there will be a 3rd order IM product at a frequency of p+r-q. This product must be at least 55dB down compared with the 0dB reference level. The frequency q must be varied between p and r such that the IM product mentioned above is maximum.
It can be understood that such a performance can only be realized with a class-A amplifier.

Some experimental devices in SOT 119 encapsulation have been tested in this way at f= 225MHz.
The DC adjustment was:
V_{DS}=28V; I_D= 2.85A.
The output matching network was aligned by means of a dummy load consisting of the parallel connection of a 6 Ohm resistor and a 150pF capacitor. This combination gave optimum results concerning IM distortion.

At a heatsink temperature of $25^{o}C$ the average output power was 17.3W peak sync at d_3= -55dB and the average power gain was 16.3dB.

When the heatsink temperature was increased to $70^{o}C$ the output power dropped to 16.3W and the gain to 15.9dB.

This performance is comparable to that of the bipolar type BLV 33F although the output power is somewhat lower and the power gain somewhat higher.

Investigation in this area will be continued.

ACKNOWLEDGMENT

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J.v.Uden and F.Wilbers: DC incl.temperature measurements.
L.v.d.Burgt: thermal investigations,
J.v.d.Giesen: S-parameter measurements,
J.Gajadharsing and G.Lukkassen: Test and application circuits and RF measurements.

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APPLICATION

Report no.: RNR-1-498-1986-AS / NCO 8602
Author : G.Lukkassen
Date : 1986-09-17

A WIDEBAND POWER AMPLIFIER (25-110 MHz) WITH THE MOS TRANSISTOR BLF 245

S U M M A R Y

For military communication purposes a wideband class-AB power amplifier has been designed around the BLF 245 with the frequency range 25 to 110 MHz.

The DC-setting is V_D=28V and I_{DQ}= 200mA.

In the input and output matching networks asymmetrical 1:4 transformers on 4C6 ferrite core material have been applied.

The main properties are:

gain at P_o= 27.5W	: 17.7 +/- 0.5dB
bandwidth	: 25 - 110 MHz
V_D	: 28V
I_{DQ}	: 200mA
efficiency	: 55-67%
input VSWR	: $\leq$ 1.6

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APPENDIX 20

1. INTRODUCTION

The BLF 245 is an RF power MOS transistor for the VHF frequency range in a SOT 123 encapsulation.
For application in military communication equipment a wideband power amplifier has been developed with a frequency range from 25 to 110MHz.
The transistor operates in class-AB at V_{DS}=28V and a quiescent current I_{DQ}=200mA. The useful outputpower is in the range of 25-30W.

2. DESIGN OF THE AMPLIFIER

2.1. General remarks

The amplifier has been developed with 1:4 impedance transformers in the input as well as in the output circuit. These transformers of the transmission line type with a ferrite core transform the 50 Ohm system impedance at the input and output to about 12.5 Ohm. An LC compensation circuit has been applied to transform this 12.5 Ohm to the optimum load impedance of the transistor. At the input a circuit matches the 12.5 Ohm to the gate impedance of the transistor and also takes care of a flat gain over the whole bandwidth.

2.2. Output circuit

For an optimum alignment of the output circuit the transistor has been replaced by a dummy. This dummy consists of a resistor of 12 Ohm parallel with a capacitor of 82pF.
The real part of the dummy has been determined by the available drain voltage and the required output power
$R_L = V_D^2/2P_o \rightarrow R_L = 28^2/2.30 = 13.1$ Ohm. This is near to the value of 12.5 Ohm mentioned in section 2.1.
The capacitor is about 15% higher than the output capacitance of the transistor.
The RF choke at the drain side must have a sufficient high reactance at the lower end of the frequency range.
Choosing this reactance appr. a factor 5 higher than the transistor loadresistance we get an inductance of 455nH for L_4.
The output capacitance of the transistor can be compensated according to the Appendix.
The result is: L_6=18.6nH and C_{11}=82pF.

To transform the achieved 12.5 Ohm to the 50 Ohm system impedance an asymmetrical 1:4 transformer has been used.
Information about this kind of transformation can be found in Ref.1 and Ref.2.
For the transformer a toroid of 4C6 material has been used.
Dimensions: 23x14x7mm. On this toroid 5 turns of two 0.7mm twisted enamelled Cu-wires are uniformly distributed and connected as shown in Fig.1

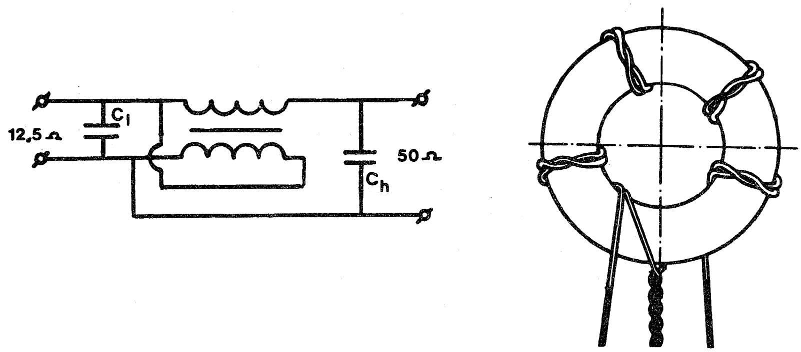

Fig.1 Output transformer

With the aid of a network analyser the transformer has been corrected for higher frequencies.
With C_l=68pF and C_h=12pF the return losses in the range 20-140MHz are better than -30dB (VSWR<1.07).
Optimization of the complete output circuit has been carried out by measuring the return losses at the output with the network analyser under swept condition (see Fig.2).

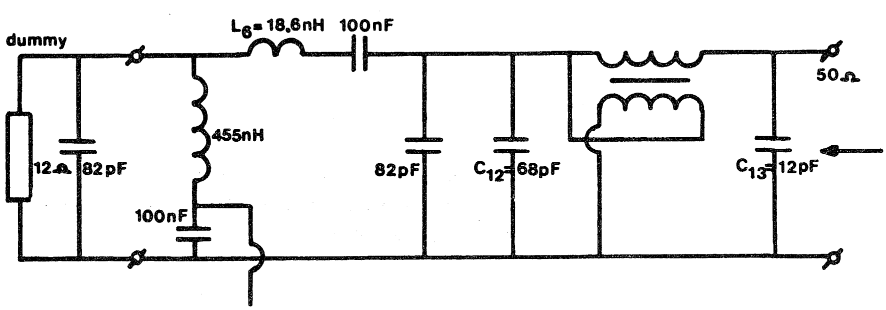

Fig.2 Output circuit before optimization

Fig.3 on page 9 shows the return losses of the output circuit before and after practical optimization. By decreasing L_6 to 10nH and C_{12} to 43pF the return losses improved about 10dB in the frequency range 20 to 140MHz to -20dB (VSWR=1.22).

2.3. Input circuit

As mentioned in section 2.1. a special circuit matches the input impedance of the transistor to 12.5 Ohm and also takes care of a sufficient flat gain over the whole bandwidth.
To determine the gate-source impedance and the gain of the transistor in combination with the output circuit described in section 2.2., narrow band input circuits have been used at several frequencies. By tuning such an auxiliary input circuit the gain of the transistor in combination with the output circuit can be measured directly.
In case the input circuit has been tuned the output impedance of this circuit is the conjugate complex of the input impedance of the transistor.
Fig.4 to 6 on page 10 and 11 give the input impedance and the gain of the transistor in combination with the output circuit. The matching network chosen at the input of the transistor is depicted inFig.7.

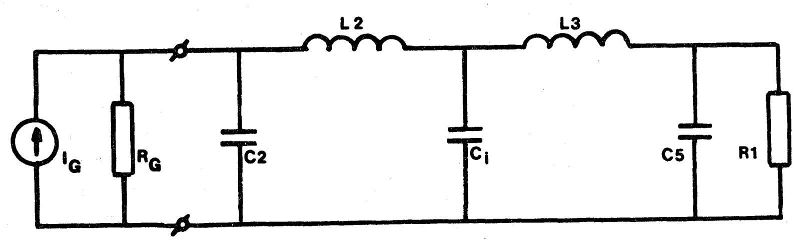

Fig.7 Input matching circuit

Ci represents the input capacitance of the BLF 245 which is appr.220pF (see Fig.5). Across this capacitor a constant voltage versus frequency from 25 up to 110MHz has to be developed. Provided Ci is an ideal capacitance the optimum dimensioning of this network is as follows:

$R_G = R_1 = 1.6/(\omega_c . C_i) = 10.5$ Ohm

$C_2 = C_5 = 0.386\ C_i = 85$pF

$L_2 = L_3 = 0.997\ R_1/\omega_c = 15.1$nH

in which ω_c is the maximum angular frequency. The calculated voltage variation across C_i is +/- 0.36dB and the maximum VSWR seen by the generator is 1.36.

Deviating from this calculation, for the ease of transformation, R_G and R_1 have been chosen 12.5 Ohm. Further the resistive component of C_i is substantial especially at higher frequencies.

Therefore the values of the components have been changed in a computer optimization program for a maximally flat gain and a low input VSWR.

This optimization results in a gain of 17.5dB with a variation of +/- 0.17dB and a maximum VSWR=1.177. These results have been achieved by changing the components of Fig.7:

C_2=97pF, C_5=102pF, L_2=17.6nH, L_3=29nH and R_1=12 Ohm.

The remaining part of the transformation from 12.5 Ohm to the 50 Ohm system impedance has been accomplished with a transformer similar to the output transformer. However the input transformer has been wound on a core consisting of 2 small toroids of 4C6 material (6x4x2mm).

On this core 6 turns of two 0.25mm twisted enamelled Cu-wires are uniformely distributed similar to the output transformer described in section 2.2. (See Fig.1 on page 4).

With correction capacitors at the high ohmic and the low ohmic side of respectively 8.2pF and 47pF the return losses in the range 20-140MHz are better than -27dB (VSWR$\leqslant$1.1).

For the practical optimization of the complete inputcircuit the transistor has been adjusted at V_D=28V and a quiescent current I_{DQ}=200mA. The gain and input return losses have been measured in the frequency range of 20 up to 110MHz.

The best results have been achieved by changing the correction capacitor C_3 from 47pF to 62pF and by executing R_1 as a parallel connection of 5 resistors of 61.9 Ohm.

Fig.8 on page 12 gives the complete circuit diagram of the BLF 245 wide band amplifier and Fig.9 on page 13 gives the corresponding parts list.

3. MEASURED PERFORMANCE

3.1. Constant inputpower

Fig.10 to 12 on page 14 and 15 give the gain, efficiency and outputpower versus the frequency at a constant inputpower (P_i=0.5W).

In the frequency range of 25MHz to 110MHz the gain is 17.2 to 17.9dB, the efficiency 55 to 70% and the output power 26.5 to 30.5W.

3.2. Constant output power

Fig.13 and 14 on page 16 give the gain and efficiency versus the frequency at a constant output power (P_o=27.5W) and heatsink temperatures of 25^{o}C and 70^{o}C.

Fig.15 and 16 on page 17 give the input return losses and the 2e and 3e harmonics of the output signal also versus the frequency. The return losses have been measured at a heatsink temperature of 25^{o}C and 70^{o}C. The harmonics have been measured at 25^{o}C. By increasing the heatsink temperature from 25^{o}C to 70^{o}C the gain decreases about 1.2dB. The heatsink temperature has no influence on efficiency and return losses.

At 25^{o}C the gain of the amplifier varies from 17.2 to 18.2dB, the efficiency from 55 to 67% and the return losses at the input are at least -14dB (VSWR$\leqslant$1.6).

Also the 2e and 3e harmonics are at least 14dB down.

3.3. Constant frequency

Fig.17 to 19 on page 18 and 19 give the output power versus input power and the gain and efficiency versus output power at 4 frequencies.

3.4. Stability

Applying an R&S PTU low pass filter at the output of the amplifier stability measurements have been carried out. Choosing a low pass frequency as close as possible above the measuring frequency the amplifier was stable through the whole frequency range of 25 to 110MHz.

3.5. Mismatch

The amplifier has been tested for load mismatch at all phase angles. Up to VSWR=10:1 the amplifier is stable. At VSWR=20:1 the amplifier is only stable below 70MHz. However also at higher frequencies degradation of the RF perfromance did not occur.

4. CONCLUSIONS

Based on the results presented in this report it may be concluded that it is quite possible to design a wideband amplifier from 25 to 110MHz with a very good performance using the MOS transistor BLF 245.
The main properties are:

- Bandwidth : 25-110MHz
- V_D : 28V
- I_{DQ} : 200mA
- Gain (P_o = 27,5W) : 17.7 +/- 0.5dB
- Efficiency : 55-67%
- Input VSWR : $\leqslant$1.6

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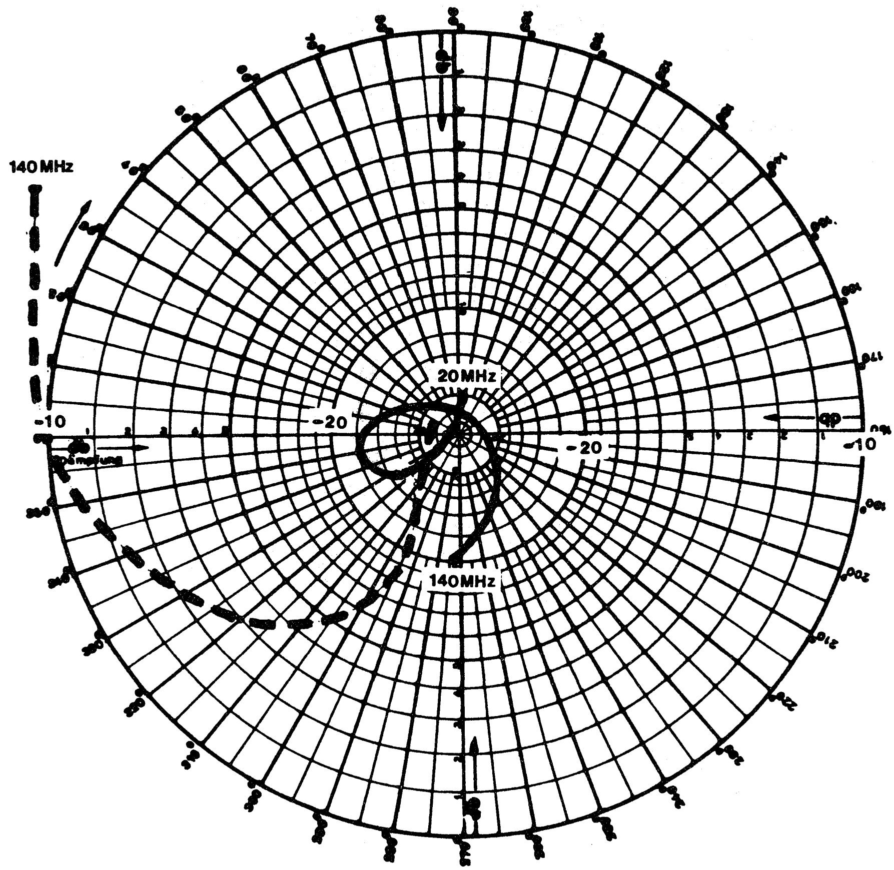
Application report ECO 7703
Power Transformers for the Frequency Range 30-80MHz.

G.Lukkassen



Before optimization

After optimization

Fig.3 Return losses output circuit

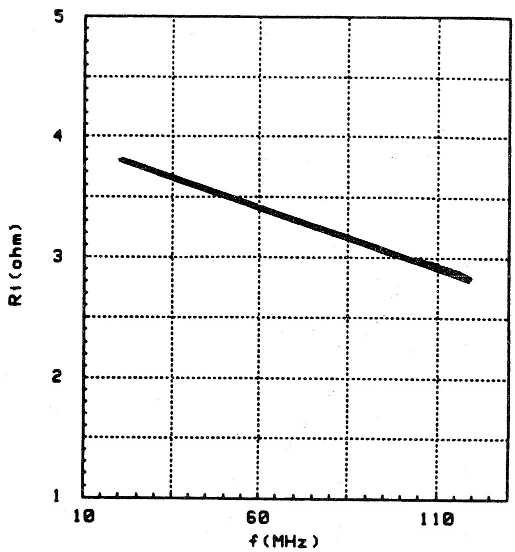


Fig.4 Real part of input impedance of loaded transistor

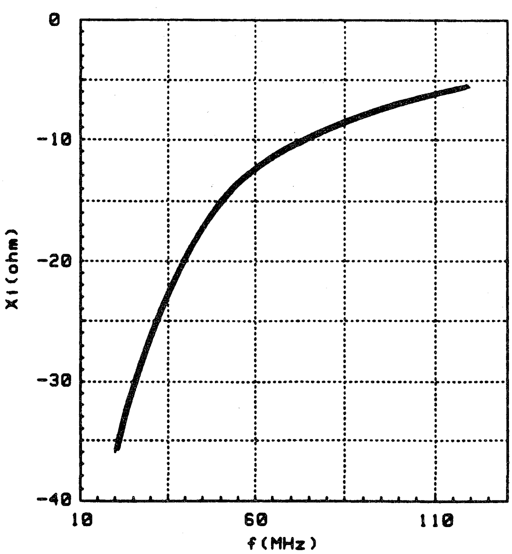


Fig.5 Imaginary part of input impedance of loaded transistor

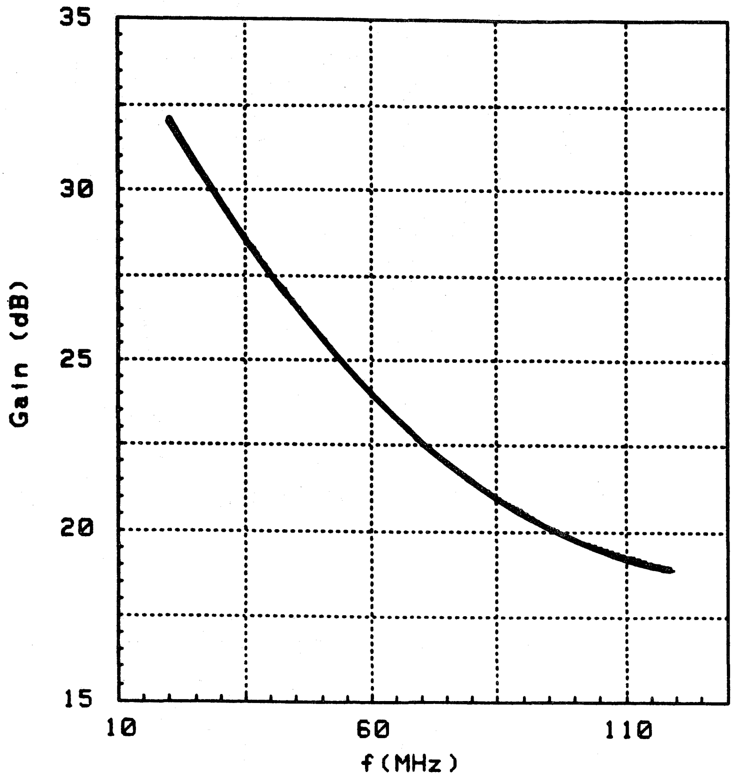


Fig.6 Gain of loaded transistor

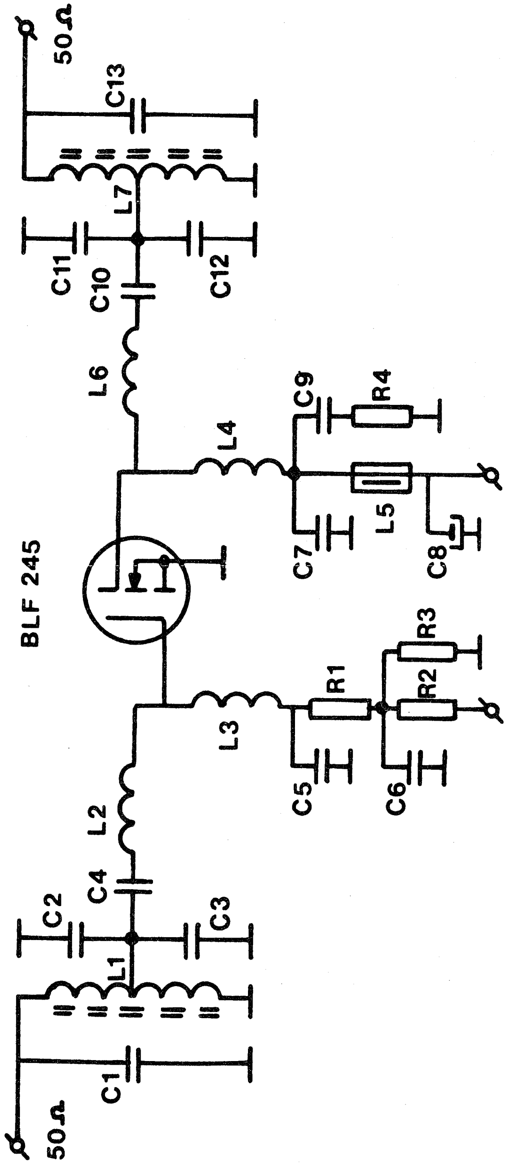


Fig.8 Circuit diagram of the BLF 245 wide band amplifier

WIDEBAND POWER AMPLIFIER WITH BLF 245 (f=25-110 MHz)

C1 =	8.2pF	multilayer ceramic chip capacitor *
C2-C5 =	100pF	multilayer ceramic chip capacitor *
C3 =	62 pF	multilayer ceramic chip capacitor *
C4=C10=	10nF	multilayer ceramic chip capacitor (cat.nr.2222 852 47103)
C6=C7=	100nF	multilayer ceramic chip capacitor (cat.nr.2222 852 47104)
C8=	2.2uF	electrolytic capacitor
C9=	3x100nF	multilayer ceramic chip capacitor(cat.nr.2222 852 47104)
C11=	82pF	multilayer ceramic chip capacitor *
C12=	43pF	multilayer ceramic chip capacitor *
C13=	12pF	multilayer ceramic chip capacitor *

L1=	2x Ferroxcube toroids, grade 4C6 (6x4x2mm)(cat.nr.4322 020 97160) with 6 turns of 2x0.25mm twisted enamelled Cu-wire (See fig. 1)
L2=	17.6nH, 2 turns enamelled Cu-wire (0.6mm) int.dia.:3mm ,length 2.5mm, leads 2x5mm
L3=	28.8nH, 3 turns enamelled Cu-wire (0.6mm) int.dia.:3mm ,length 3.2mm, leads 2x5mm
L4=	455nH, 12 turns enamelled Cu-wire (1mm) int.dia.:7mm ,length 16.5mm, leads 2x5mm
L5=	Ferroxcube h.f.choke,grade 3B (cat.nr.4312 020 36642)
L6=	10nH,1 turn enamelled Cu-wire (1mm) int.dia.:3mm leads 2x3mm
L7=	Ferroxcube toroid,grade 4C6 (23x14x7mm) (cat.nr.4322 020 97190) with 5 turns of 2x0.7mm twisted enamelled Cu-wire (See fig. 1)

R1=	12.4 Ohm,parallel connection of 5 metal film resistors 61.9 Ohm (cat.nr.2322 151 76199)
R2=	1K Ohm, metal film resistor (cat.nr. 2322 151 71002)
R3=	1M Ohm, metal film resistor (cat.nr. 2322 151 71005)
R4=	10 Ohm, metal film resistor (cat.nr. 2322 153 51009)

PC-board: double Cu-clad, 1.6mm epoxy fibre-glass (ε_r=4.5)

* American Technical Ceramics type 100B or capacitor of same quality.

Fig.9 Parts list of the BLF 245 wideband amplifier

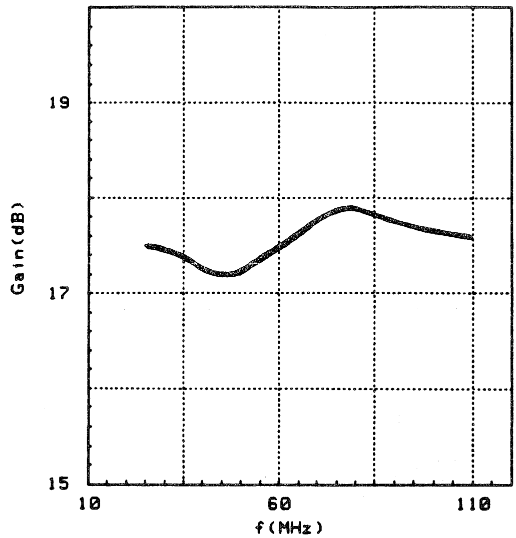


Fig.10 Gain at P_i=0.5W

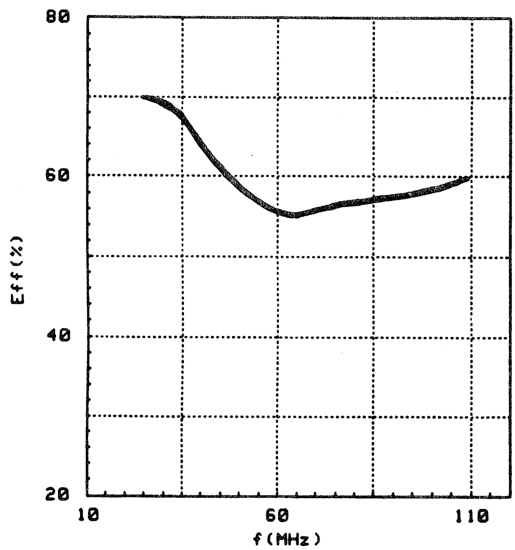


Fig.11 Efficiency at P_i=0.5W

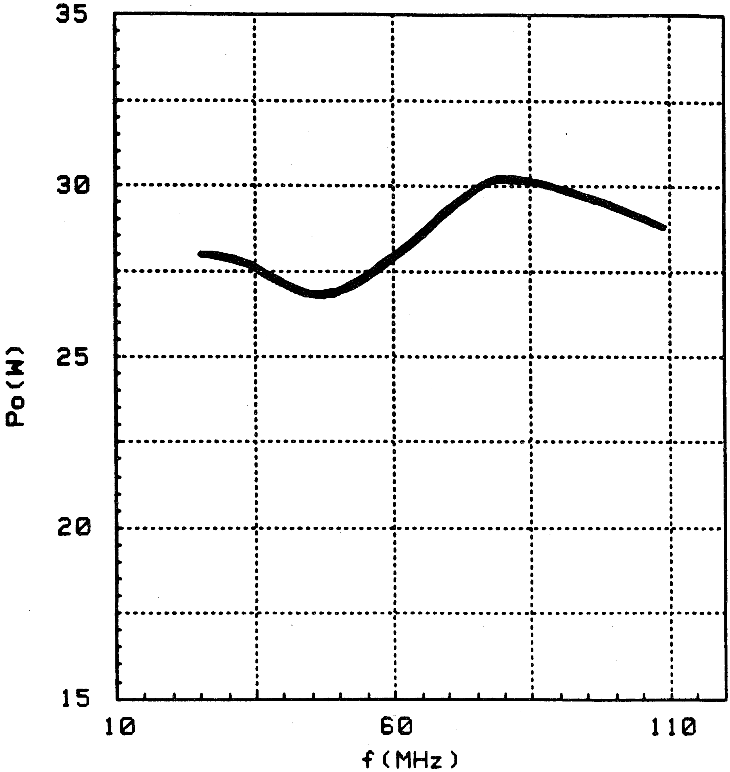


Fig.12 Output power at P_i=0.5W

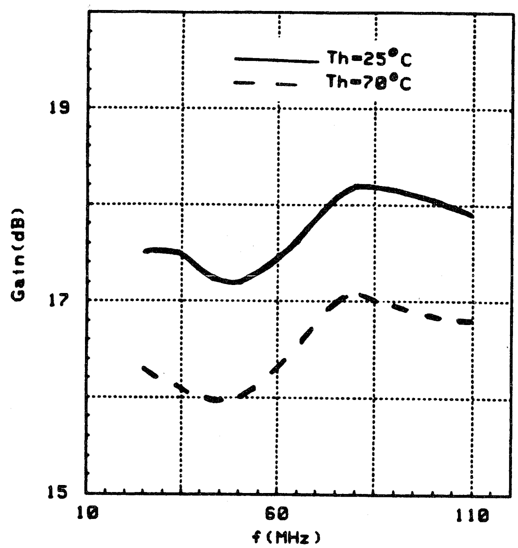


Fig.13 Gain at P_o=27.5W

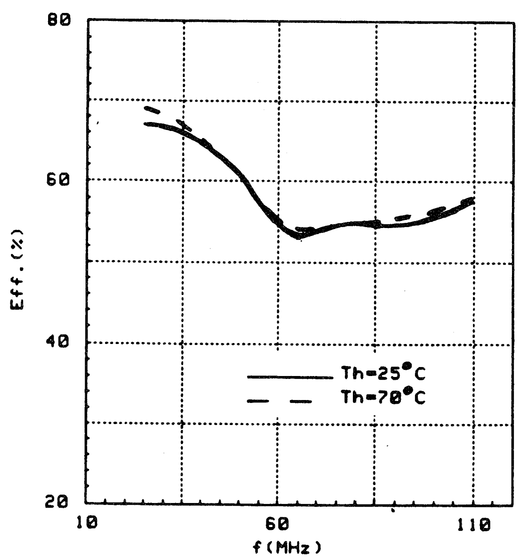


Fig.14 Efficiency at P_o=27.5W

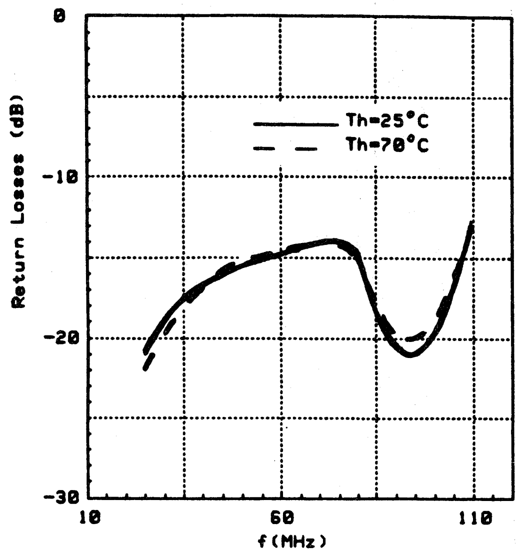


Fig.15 Input return losses at P_o=27.5W

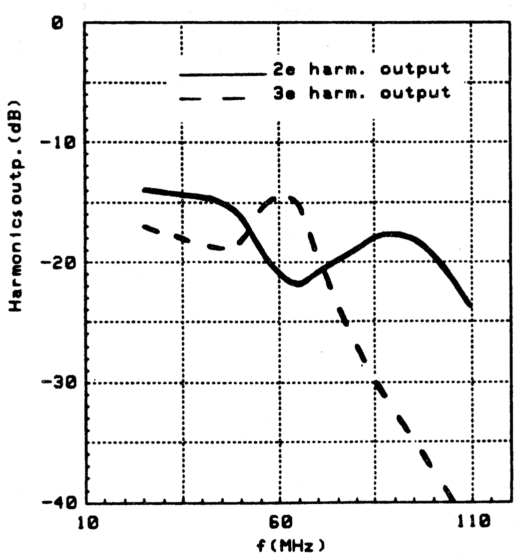


Fig.16 Output harmonics at P_o= 27.5W

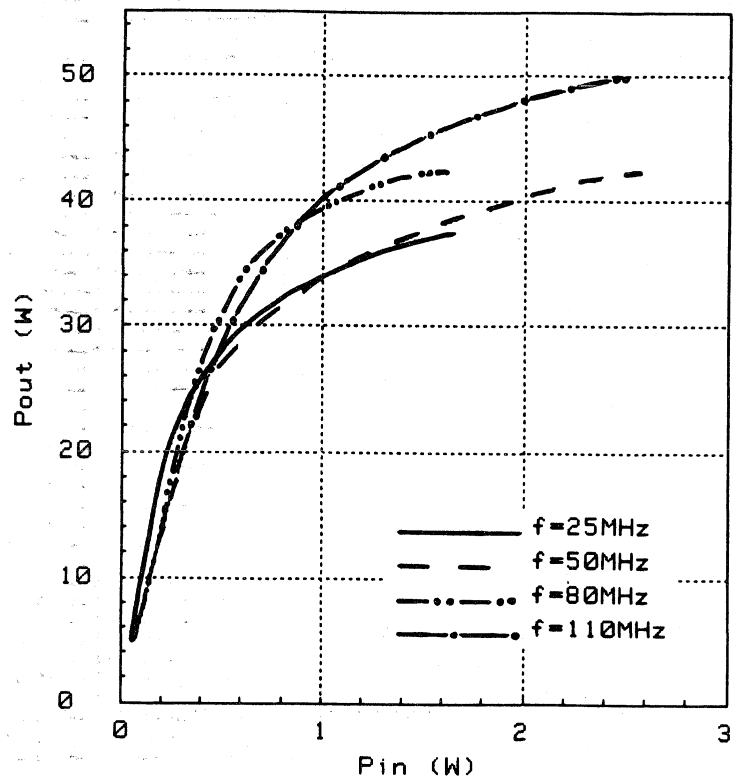


Fig.17 Output power versus input power

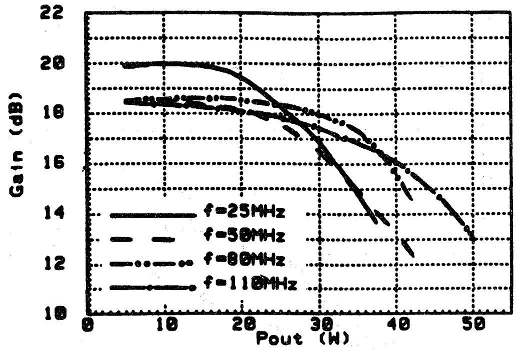


Fig.18 Gain versus output power

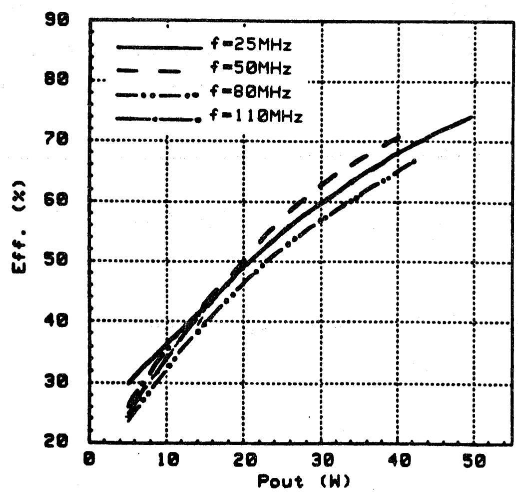
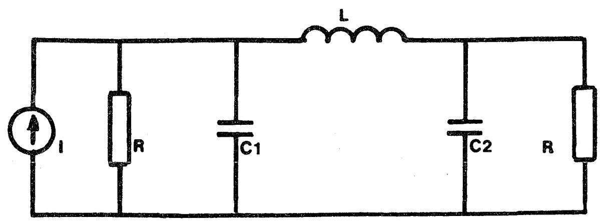


Fig.19 Efficiency versus output power

APPENDIX

The output capacitance of a transistor can be compensated over a certain bandwidth by absorbing it in a low-pass Chebyshev π-section.



If C1 is the transistor output capacitance the components L and C2 must be added.

$C2=C1=C$

The normalized value of C is:

$A=\omega_m CR$

in which $\omega_m = 2\pi f_{max}$

Now we can calculate the normalized value of L with:

$B=8A/(3A^2+4)$

where $B=\omega_m L/R$

The maximum VSWR of this network can be calculated with the following procedure.

1. Determine $\gamma=1/A$
2. $X=\gamma+\sqrt{\gamma^2+1}$
3. $VSWR=\{(X^3+1)/(X^3-1)\}^2$

In our amplifier:

R= 12.5 Ohms

C=82pF

This gives:

A=0.784

B=1.029

L=18.62nH

γ=1.412

X=3.142

VSWR=1.138

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Date : 1987-03-11

TITLE

A WIDEBAND 30 WATT PUSH-PULL AMPLIFIER FOR THE FREQUENCY RANGE 25-110MHz WITH TWO MOS TRANSISTORS BLF244 (V_{DS}=28V)

SUMMARY

In this report a description is given of a wideband power amplifier for the frequency range 25-110MHz. It utilizes two BLF244 MOS transistors in a push-pull configuration. The transistors are adjusted in class-B with a quiescent drain current of 50mA each at V_{DS}=28V. The main properties at P_{out}=30W are:

Powergain	=	15.7 ± 0.7dB
Drain efficiency	=	60.6 ± 3.3%
Input VSWR		≤ 1.5
Second harmonic level		≤ -40dB

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1. INTRODUCTION

A wideband push-pull power amplifier has been developed for the frequency range 25-110MHz. The design is based on the BLF244, a silicon N-channel enhancement mode vertical D-MOS transistor designed for large-signal amplifier applications in the VHF range. This device can deliver 15W output power at 175MHz when operated from a 28V supply. The transistor has a 4-lead flange envelope with a ceramic cap (SOT123).
The objective was to design and construct a 30W wideband amplifier with high gain and efficiency and low input VSWR and second order distortion. With respect to gain and efficiency a reasonable flatness was desired. The push-pull design is employed because of its low second order distortion.
The design and practical realization of this amplifier are described in the following chapters.

2. AMPLIFIER DESIGN

2.1. General

The schematic set up of the amplifier is depicted in Fig.1.

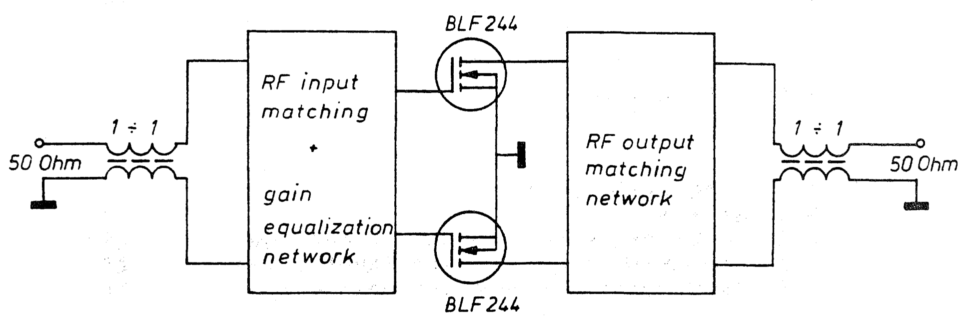

Fig.1.

Two 1÷1 balance to unbalance transformers are applied; one for splitting the single-ended input source into two out of phase sources driving the transistor-inputs, the other for adding the outputs from the transistors.
Transmission line transformers are employed because of there excellent broadband response. These transformers consist of a twisted-wire-pair transmission line wound on a ferrite toroid.

At the input side a special matching network is applied to obtain a low VSWR and compensation for variation in gain with frequency.
The matching network at the output side provides the transistors with the optimum load for an output power of 30W at V_{DS}=28V.

2.2. Powergain, input- and output impedance

The design has been started by determining powergain input impedance and output impedance of the transistor for the frequency range 25-110MHz.

First the output impedance was determined.
For HF and VHF the optimum load resistance RL can be calculated with reasonable accuracy with the formula:

$$R_L = \frac{V_{DS}^2}{2 \cdot P_o} \qquad (1)$$

For V_{DS}=28V and P_o=15W we get: R_L=26.1Ω.
The output impedance is the parallel combination of the output capacitance C_o of the transistor and the optimum load resistance.
Because of the large drain voltage swing the effective output capacitance C_o is approx. 15% higher than the value of C_{oss}. For BLF244 C_{oss} is typical 38pF, so C_o is equal to 43.7pF. So the output impedance of this transistor can be represented by 26.1Ω//43.7pF for the whole frequency range.
Second the large-signal input impedance and powergain versus frequency were determined by measurement. For this purpose a single-ended test amplifier was constructed. This amplifier was matched at the output side to a load of 25Ω by a broadband matching network. Dimensioning of this network was based on a practical dummy transistor of 25Ω//43pF. The maximum VSWR measured within the band was 1.16.
At the input side tunable narrowband matching networks were applied at several frequencies. By tuning this amplifier for minimum return loss at P_o=15W the powergain was measured directly.

For measurement of the input impedance the DC power, signal source and transistor were disconnected from the amplifier and the signal source circuit connection was terminated with 50Ω. After that the impedance was measured at the gate connection of the transistor. The input impedance of the transistor is the conjugate of the measured impedance if the circuit doesn't contain resistive components.

This procedure was repeated at several frequencies in the band to get sufficient data for the design. Fig.2,3 and 4 on page 23 present the data in graphical form.

2.3. Output matching section

Because of the symmetrical set-up of this amplifier its matching sections can be divided into two equal parts. Each part belonging to one transistor. In the next discussion one half of the output matching section will be considered.

As mentioned in the previous section the optimum load resistance for P_o=15W and V_{DS}=28V is 26.1Ω according to equation (1).
When two of these transistors are used in a push-pull configuration the optimum load resistance adds up to 52.2Ω. This value is very close to 50Ω to which these transistors have to be matched. So, if we choose the optimum load resistance to be 50Ω we can suffice with a 1÷1 balance to unbalance transformer.
The output capacitance C_o of the transistor can be compensated over a certain bandwith by absorbing it in a low-pass Chebyshev π-section, see Fig.5.

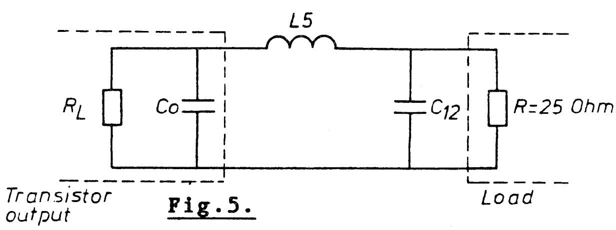

Fig.5.

R_L represents the optimum load resistance for the transistor. The components L5 and C12 can be determined with the following formulae if R_L=R and-C_o=C12=C:
The normalized value of C is:

$$A = \omega_m \cdot C \cdot R \qquad (2)$$

in which $\omega_m = 2 \cdot \pi \cdot f_{max}$

The normalized value of L5 can be calculated as follows:

$$B = \frac{\omega_m \cdot L5}{R} \qquad (3)$$

$$B = \frac{8 \cdot A}{3 \cdot A^2 + 4} \qquad (4)$$

The maximum VSWR of this network follows from:

$$VSWR_{max} = \left\{ \frac{x^3 + 1}{x^3 - 1} \right\}^2 \qquad (5)$$

in which $x = \gamma + (\gamma^2 + 1)^{1/2}$

and $\gamma = 1/A$

For this section R=25Ω and C=43pF. This results in:
A=0.7430 → B=1.0509 → L5=38nH and $VSWR_{max}$=1.156.

In practice this circuit comprises some additional components, see Fig.6.

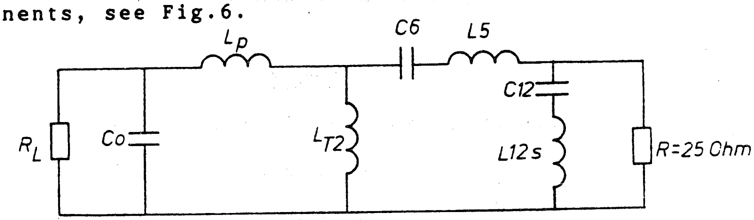

Fig.6

These are:

L_p - the parasitic drain and source inductance which has been accounted for by this way. Its value is approx. 1.4nH.

L_{T2} - the drain choke inductance which equals approx. 0.8μH. Determination of this inductance will be treated in a later chapter.

L_{12S} - the parasitic series inductance of C12, which is approx. 1nH.

C_6 - the DC-blocking capacitor which is also employed for low frequency compensation of L_{T2}.

The value of C_6 is calculated with the aid of the information given in ref.[1]. The drain load circuit for low frequency is shown in Fig.7.

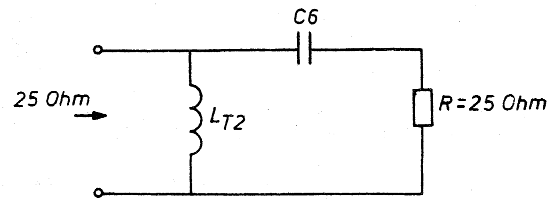

Fig.7

Compensation according to ref [1] gives at f=25MHz:

C_6=1.28nF with $VSWR_{max}$ = 1.04.

The circuit in Fig.6 was optimized for the frequency range 25-110MHz. For this purpose a computer optimization program was used. The criterion used was for overall minimum VSWR with respect to 25Ω.

The results before and after optimization are shown in the table below.

Before optimization		After optimization	
C_6	= 1.28nF	C_6	= 7.9nF
L_5	= 38nH	L_5	=34.6nH
C_{12}	= 43pF	C_{12}	=37.9pF
$VSWR_{max}$	= 1.169	$VSWR_{max}$	= 1.098

2.4. Input matching section

The purpose of the input matching section is twofold. First to match the transistor input impedance to the source impedance of 50Ω with a sufficiently low VSWR across the frecuency band.
Second to compensate the variation in gain with frequency.

The input matching section chosen is depicted in Fig.8 for one transistor.

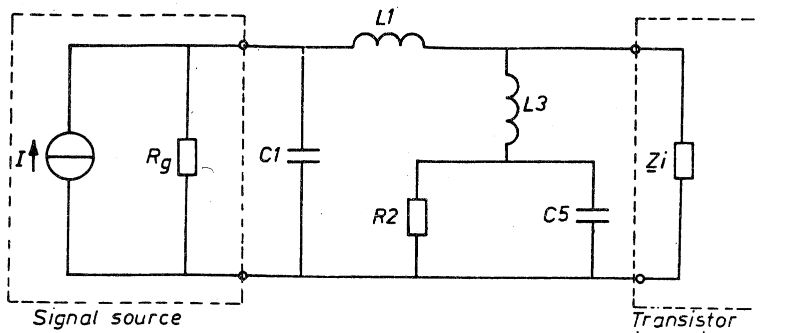

Fig.8

Since the input impedance of the transistor is strongly capacitive, Zi can be approximated by an ideal capacitor. This network can then be treated as a symmetrical double pi-section, see Fig.9.

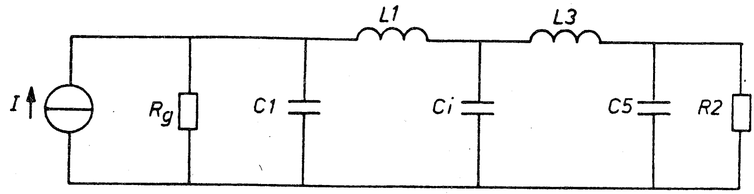

Fig.9

In order to get sufficient gain flatness a constant voltage has to be developed across capacitor Ci. For optimum dimensioning of this network the following formulae are valid:

$$Rg = R2 = 1.6/(\omega_m \cdot Ci) \tag{6}$$

$$C1 = C5 = 0.386 \times Ci \tag{7}$$

$$L1 = L3 = 0.997 \times Rg/\omega_m \tag{8}$$

in which ωm is the maximum angular frequency.
These formulae have been obtained by a computer optimization program which also indicates that the maximum voltage variation across Ci is ± 0.36dB and the maximum VSWR seen by the generator 1.36. When the input capacitance at the lowest frequency is chosen, which is approx. 117pF, we find that:

Rg = R2 = 19.8Ω
C1 = C5 = 45.2pF
L1 = L3 = 28.6nH

In practice Rg is equal to 25Ω and Zi varies with frequency. This required a re-optimization of this network with the actual values of Zi and Rg. The values calculated above were used as the initial values and parasitics of C1 and C5 were included. The target gain was set to 17.5dB.
The results of this optimization are shown in the table below.

Before optimization		After optimization	
C1	= 45.2pF	C1	= 60.1pF
C5	= 45.2pF	C5	= 47.5pF
L1	= 28.6nH	L1	= 36 nH
L3	= 28.6nH	L3	= 43.8nH
R2	= 19.8Ω	R2	= 20.8Ω
Rg	= 25 Ω	Rg	= 25 Ω
$VSWR_{max}$	= 1.812	$VSWR_{max}$	= 1.376
$G_p min$	= 15.8dB	$G_p min$	= 17.1dB
$G_p max$	= 16.9dB	$G_p max$	= 17.9dB

3. TRANSFORMER DESIGN

3.1. General

As mentioned before transformers employed at the input and output side utilize twisted-wire-pair transmission lines wound on a toroidal core.
The windings are uniformly distributed around the toroid. The required characteristic impedance of the transmission lines is 50Ω. In practice Zo will differ from this required value and compensation measures will be necessary [2]. This can be achieved with:
- parallel capacitances across input and output terminals of the transformers if Zo>50Ω.
- inductances in series with the input and output terminals of the transformer if Zo<50Ω.

The result of this compensation is an exact match at the maximum frequency. There will be however, a slight mismatch at low frequency which is many times smaller then that at the maximum frequency without compensation. Because the amount of HF compensation will depend on the circuit layout and the exact transformer construction no calculations will be made on this aspect of the transformers. The amount of compensation will be determined in the circuit by employing adjustable capacitors.

3.2. Design of the output transformer

The characteristic impedance of 50Ω for the transmission line of the output transformer has been obtained with enamelled copper wire of 0.6mm diameter. Its diameter with isolation included is 0.66mm.
The number of twists applied are 2 per centimeter.

A suitable core material for this frequency range is Philips 4C6 grade available in several sizes of toroid. The size of the toroid is determined by the maximum allowable dissipation which is limited to $350mW/cm^3$ to prevent excessive rise in temperature. Designing for a maximum of 1% power loss in the core (300mW) the minimum effective volume required is:

$$Ve_{min} = \frac{P_{loss}}{350\ mW/cm^3} = \frac{300}{350} = 0.85\ cm^3$$

The smallest toroid that is suitable is a core with dimensions Dxdxh=23x14x7mm corresponding with a effective core volume of $1.79cm^3$. As a result the core loss reduces to $170mW/cm^3$.
According to reference [3] this corresponds to a maximum flux density ($\hat{B}$) of 0.2mT at 110MHz. The required number of turns is determined by the ratio Rp/L in which Rp is the loss resistance that represents the core loss and L the inductance in parallel with the output terminals, see reference [4].
This ratio is equal to:

$$R_p/L = \frac{\omega^2 \cdot \hat{B}^2 \cdot Ve}{2 \cdot \mu_o \cdot \mu_r \cdot P_{loss}} \qquad (9)$$

which amounts to:

$$R_p/L = \frac{(2 \times \pi \times 110 \times 10^6)^2 \times (0.2 \times 10^{-3})^2 \times 1.79 \times 10^{-6}}{2 \times 4 \times \pi \times 10^{-7} \times 120 \times 0.3} = 472\ Ohm/\mu H$$

To keep the core loss below 1% we must keep the parallel loss resistance above 5000Ω with reference to 50Ω. This means an inductance of:

$$L = Rp/472 = 10.6\,\mu H$$

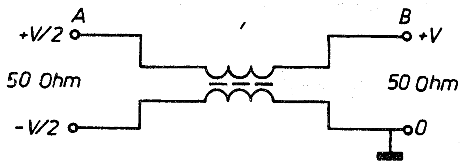

Fig.10

Between point A and B in Fig.10 the voltage is one half of the output voltage. Therefore the inductance between these points must be a quarter of that across the 50Ω terminals, so:

$$L_{AB} = L/4 = 10.6/4 = 2.65 \mu H$$

The number of turns required can be calculated with the following formula [3]:

$$L = A_L . N^2 \qquad (10)$$

$$A_L = \frac{0.4 \times \pi \times \mu_r}{\Sigma l/A} \qquad (11)$$

in which: A_L is the inductance factor in [nH]
$\Sigma l/A$ is the core constant in $[mm^{-1}]$ given in [3]
N is the number of turns
μr is the relative permeability (120 for grade 4C6).

For a toroid of 23mm the core constant is $1.81 mm^{-1}$. So, the inductance factor amounts to:

$$A_L = \frac{0.4 \times \pi \times 120}{1.81} = 83.3\, nH$$

and the required number of turns:

$$N = \sqrt{\frac{2.65 \times 10^3}{83.3}} = 5.6 \text{ turns}$$

In practice the number of turns will be 6, so the inductance in parallel with the output terminals rises to:

$$(6/5.6)^2 \times 10.6 = 12 \mu H$$

This corresponds to a reactance of 1885Ω at 25MHz which is high enough to be neglected.
The core loss reduces to:

$$(5.6/6)^2 \times 1\% = 0.87\%$$

The measured value of L_{AB} was approx. 3.5μH.

3.2. Design of the input transformer

The input transformer is of the same type as the output transformer and is also designed in the same way.

To obtain a characteristic impedance of 50Ω for the windings enamelled Cu-wire with a bare diameter of 0.50mm is used. The diameter with isolation included is 0.55mm. The number of twists applied is 2 3/4 per centimeter.

Allowing an input power level of 1.5W the minimum effective volume for 1% power loss in the core is:

$$Ve_{min} = 15/350 = 0.043\ cm^3$$

The smallest toroid that suits our need is a type with dimensions Dxdxh=9x6x3mm. The effective core volume is $0.105cm^3$, so the core loss reduces to $143mW/cm^3$. This corresponds to a maximum flux density $\hat{B}$ of approx. 0.18mT at f=110MHz according to ref.[3].
The ratio Rp/L amounts to:

$$Rp/L = \frac{(2\times\pi\times110\times10^6)^2\times(0.18\times10^{-3})^2\times0.105\times10^{-6}}{2\times4\times\pi\times10^{-7}\times120\times0.015} = 360\ Ohm/\mu H$$

For 1% loss L amounts to:

$$L = Rp/360 = 5000/360 = 13.9\,\mu H$$

So $$L_{AB} = 13.9/4 = 3.48\,\mu H$$

The required number of turns for a 9mm toroid with a core constant of $5.17mm^{-1}$ is:

$$A_L = \frac{0.4 \times \pi \times 120}{5.17} = 29.2\,nH$$

$$N = \sqrt{\frac{3.48 \times 10^3}{29.2}} = 11\ turns$$

An inductance of $L=13.9\mu H$ corresponds to a reactance of 2183Ω at 25MHz which is high enough to be neglected.

According to measurements 10 turns were sufficient to obtain $L_{AB} \simeq 3.5\mu H$.

3.3. The tapped choke (T2)

The chokes in the drain circuits are wound on a common ferrite rod. The windings are twisted together. Constructional details are shown in Fig.11. With this arrangement the dc flux components in the core cancel out and a much smaller component results. Because a rod has a open magnetic circuit saturation effects will hardly occur.
In Fig.12 the output part of the amplifier is given in a different way.

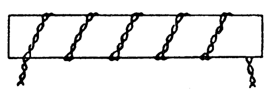

Fig.11

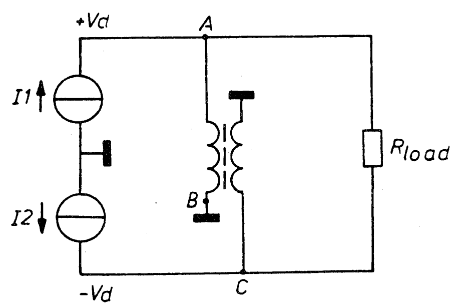

Fig.12

The current sources have a frequency spectrum in which the even order components are in phase and the odd order ones in anti-phase.
For the even harmonics the impedance between point A and C will depend on the coupling factor K between the windings with:

$$\omega L_{AB}(1 - K) \qquad (12)$$

If the coupling factor amounts to 1 points A and C will be short circuited. If the current components are in anti-phase the total inductance between these points shunts the load resistance. Because the voltage between point A and B is equal to ½ V_{AC} the total inductance L_{AC} is equal to 4 L_{AB} if the coupling factor is 1.
The reactance of this shunting inductance is allowed to be at least 4 times the load resistance or 200Ω at 25MHz. So, L_{AC} amounts to 1.27μH and L_{AB} to 0.318μH.

To obtain the inductance L_{AB} a ferrite rod grade 4B1 has been used with a length of 30mm and a diameter of 5mm. According to ref.[3] its relative permeability is equal to 20. The number of turns can be determined with:

$$N = \sqrt{\frac{L \cdot l}{\mu_o \cdot \mu_r \cdot A}} \qquad (13)$$

For L_{AB} this amounts to:

$$N = \sqrt{\frac{0.318 \times 10^{-6} \times 30 \times 10^{-3}}{4 \times \pi \times 10^{-7} \times 1/4 \times \pi \times (5 \times 10^{-3})^2}} = 4.4 \text{ turns}$$

In practice 5 turns will be used so L_{AC} will be equal to 1.6μH. The measured value for L_{AB} was 0.48μH at 25MHz. The windings are constructed of enamelled copper wire of 0.8mm diameter.

4. AMPLIFIER CONSTRUCTION

4.1. Printed circuit board and component layout

The printed circuit board of this amplifier is made of two-sided copper clad epoxy fibre glass (εr=4.5) laminate of 1/16" thickness.
Circuit components are situated on one side of this board, the other side serves as ground plane. A full sized pattern of the printed circuit board and component layout is given in Fig.14 and 15 on page 22. The parasitic inductances of the printed tracks are absorbed in the inductances of the matching networks. Connections to the ground plane are made by means of tubular rivets, straps under the source leads and at the N-connectors and the mounting screws.

4.2. Heatsink

The printed circuit board is attached to a solid copper plate, with dimensions 120x100x10mm, which functions as a heatsink.It is provided with a tube in order to control its temperature by means of a water cooling system. Good thermal contacts between transistors and heatsink is obtained by use of a heat-sinking compound.

5. AMPLIFIER ALIGNMENT

The amplifier was constructed according to the theoretical design procedures. Fig.13 on page 20 shows the total circuit diagram of this amplifier. Parallel matching components as C1,C5 and C12 are connected directly from one side of the circuit to the other. Therefore their values are exactly one half of those calculated. The component list is given on page 21.

Alignment of this amplifier was first done on a small signal basis. First the output circuit was aligned by replacing the BLF244 transistor with dummy loads, representing the conjugate of the optimum load impedance. The dummy's consisted of a 25Ω resistance and a 43pF capacitance. Several components in parallel were used to obtain symmetry and to reduce parasitic inductances. These components were soldered in an empty SOT123 header. The return loss was measured versus frequency at the load connection of the amplifier and minimized by applying compensation capacitors between the terminals of the output transformer. At the load side of the transformer 3.6pF (C13) was needed and at the transistor side C12 was increased from 20pF to 22pF. The maximum VSWR obtained was 1.22.

Alignment of the input circuit has been done with the transistors in the circuit and supply and load connected. The quiescent drain current was set to approx. 200mA per transistor and return loss was measured versus frequency. Experiments with capacitors in parallel with the input transformer terminals showed that no compensation was needed. The maximum VSWR obtained was 1.30.

After the small signal alignment the transistors were set to class-B operation by decreasing the quiescent drain current to 25mA.
The first results obtained at P_{out}=30W were:
G_p=15.6 ± 1.2dB; Eff.=61.3 ± 10%; VSWR $\leqslant$1.40 and second harmonic level $\leqslant$-33dB.

In order to improve the total performance of this amplifier especially with respect to gain flatness, variation in efficiency and second harmonic level some minor changes were introduced in the amplifier.

1. At the input side the circuit configuration shown in Fig.16A was changed into that of Fig.16B. No appreciable improvement was achieved with respect to gain performance but variations in efficiency reduced to ± 6%. However, at the lower side of the band the second harmonic level increased to -28dB.

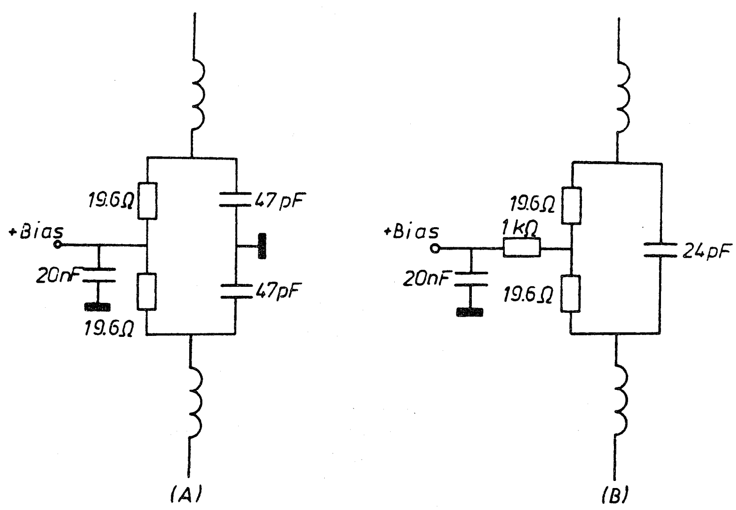

Fig.16

2. Raising the quiescent drain current to 50mA improved the gain with approx. 0.3dB. However, the efficiency decreased with approx. 0.8%.

3. A resistance of approx. 12Ω from the mid tap of the drain choke T2 to ground instead of direct grounding increased the average efficiency with approx. 1% and its variations decreased to ± 2.8%. The second harmonic level improved with 2dB.

4. The input balun was originally connected as shown in Fig.17A. For a perfect symmetrical push-pull amplifier it does not matter which terminal is grounded. In this case exchange of the terminals strongly affected the second harmonic level. For the case of Fig.17B this level improved to <-40dB.

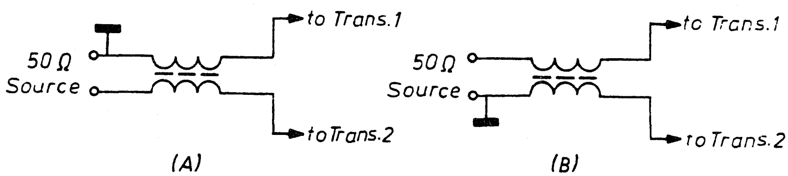

Fig.17

5. Finally the value of the resistors shown in Fig.16 was increased to 23.7Ω. This improved the gain flatness to approx. ± 0.6dB. The input VSWR increased to 1.5.

6. AMPLIFIER PERFORMANCE

6.1. General

Measurement of the amplifier performance was carried out under nominal conditions unless stated otherwise. These conditions are:

Supply voltage $V_{dd}=28V$
Quiescent drain current $I_{dq}=50mA$
Heatsink temperature $T_{hs}=25^{o}C$

Measurements were done at 10 frequencies within the band and 2 frequencies outside the band.

The BLF244 samples used, are matched on their threshold voltage V_{th}. The measured parameters of these transistors which can be relevant for balanced operation are given on the next page.

		T1	T2
V_T (V_{ds}=10V;I_d=5mA)	[V]	3.14	3.14
G_{FS} (V_{ds}=10V;I_d=750mA)	[mS]	794	838
C_{rss}(V_{ds}=28V;V_{gs}=0V;f=1MHz	[pF]	4.46	4.21

The largest asymmetry observed in the drain current was ± 3% at f=25MHz and P_o=30W.

6.2. Performance at constant output power

Measurements of the performance at a constant output power 30W were carried out at two heatsink temperatures, viz. T_h=25°C and 70°C.
The results obtained are:
Powergain = 15.7 ± 0.7dB, see Fig. 18 on page 24;
Drain eff.= 60.6 ± 3.3% , see Fig. 19 on page 24
Input return loss <-14dB (VSWR<1.50),
see Fig. 20 on page 24;
Second harmonic level <-40dB, see Fig.21 on page 25;
Third harmonic level <-14dB, see Fig.22 on page 25.

At T_h=70°C the powergain decreased with approx. 1.2dB see Fig.18 on page 24.
The other parameters showed no appreciable change.

6.3. Performance at constant input power

Performance of this amplifier was also measured at a constant input power of 700mW. The result obtained are:
Output power =27.9 ± 2.3W, see Fig. 23 on page 26;
Power gain =16.0 ± 0.3dB,see Fig. 24 on page 26;
Drain eff. =59.0 ± 4.5%, see Fig. 25 on page 26;

6.4. Performance at constant frequency

Fig 26 to 28 on page 27 show the following curves measured at 5 different frequencies:
- P_o = f (P_i)
- G_p = f (P_o)
- Eff.= f (P_o)

7. CONCLUSIONS

Using two BLF244 MOS transistors (matched on threshold voltage) in a push-pull configuration approx.16dB power gain and 60% drain efficiency have been obtained for an output power of 30W, when operated with a quiescent drain current of 50mA per transistor at V_{ds}=28V. The largest asymmetry observed in the drain current was ± 3% at P_{out}=30W and f=25MHz. The input VSWR was below 1.5.

Throughout the band the second harmonic level was lower than -40dB with reference to the fundamental. At a heatsink temperature of 70°C the powergain decreased with approximately 1dB while the other parameters showed no appreciable change.

8. REFERENCES

[1] H.Nielinger; "Optimale dimensionerung von Breitband-anpassungsnetwerken"; N.T.Z.1968, Heft 2,pp.88-91.

[2] A.H.Hilbers; "Design of HF wideband power transformers"; Philips Application information # 530,1970.

[3] Philips Data Handbook; "Ferroxcube for power, Audio/Video and accelerators"; Book C5 1986, pp.42,317 and 320.

[4] A.H.Hilbers; "Power transformers for the frequency range of 30-80MHz"; Laboratory report EC07703,1977.

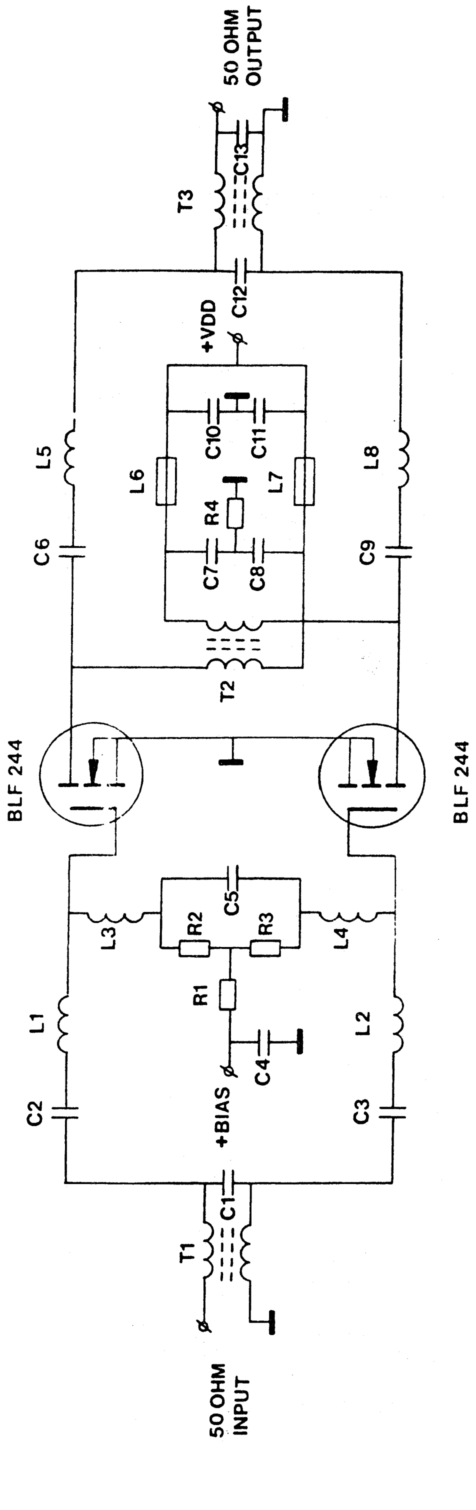

Fig.13 Circuit daigram of the wideband push-pull amplifier

LIST OF COMPONENTS

Capacitors

C1=30pF; multilayer ceramic chip capacitor *
C2=C3=C6=C7=C8=C9=10nF; multilayer ceramic capacitor (cat.nr. 2222 852 47103)
C4=2x10nF; multilayer ceramic chip capacitor (cat.nr. 2222 852 47103)
C5=24pF; multilayer ceramic chip capacitor *
C10=C11=100nF; multilayer ceramic chip capacitor (cat.nr. 2222 852 47104)
C12=22pF; multilayer ceramic chip capacitor *
C13=3.6pF; multilayer ceramic chip capacitor *

Inductors

L1=L2=27nH; 3 turns enamelled Cu-wire (0.8mm); int.dia.=4.0mm; length=6.1mm; leads 2x3.0mm
L3=L4=48nH; 4 turns enamelled Cu-wire (0.8mm); int.dia.=4.0mm; length=6.2mm; leads 2x1.0mm
L5=L8=30nH; 3 turns enamelled Cu-wire (0.8mm); int.dia.=4.0mm; length=4.8mm; leads 2x2.0mm
L6=L7= Ferroxcube RF choke, grade 3B (cat.nr.4312 020 36640)

Resistors

R1=1 KΩ; metal film resistor; 0.4W
R2=R3=23.7Ω; metal film resistor; 0.4W
R4=12.1Ω; metal film resistor; 0.4W

Transformers

T1- 1÷1 Balun; 10 turns of twisted pair of 0.5mm enamelled Cu-wire (2 3/4 twists per cm) wound on a toroidal core grade 4C6, dimensions (9x6x3)mm (cat.nr.4322 020 97191).
T2- Drain choke; 5 turns of twisted pair of enamelled Cu-wire (4.5 twists per cm) wound on a ferroxcube rod grade 4B1, dimensions (5x30)mm,
T3- 1÷1 Balun; 6 turns of twisted pair of 0.6mm enamelled Cu-wire (2 twists per cm) wound on a toroidal core grade 4C6, dimensions (23x14x7)mm (cat.nr.4322 020 97171)

Printed circuit board: double sided Cu-clad epoxy fibreglass laminate (εr=4.5), thickness 1/16"

* American Technical Ceramics capacitor type 100B.

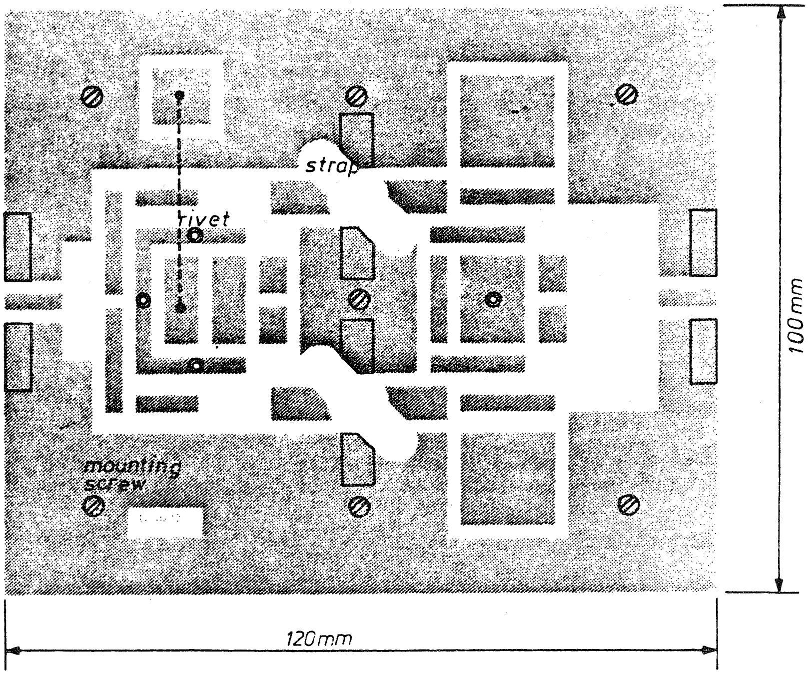

Fig.14 Printed circuit board

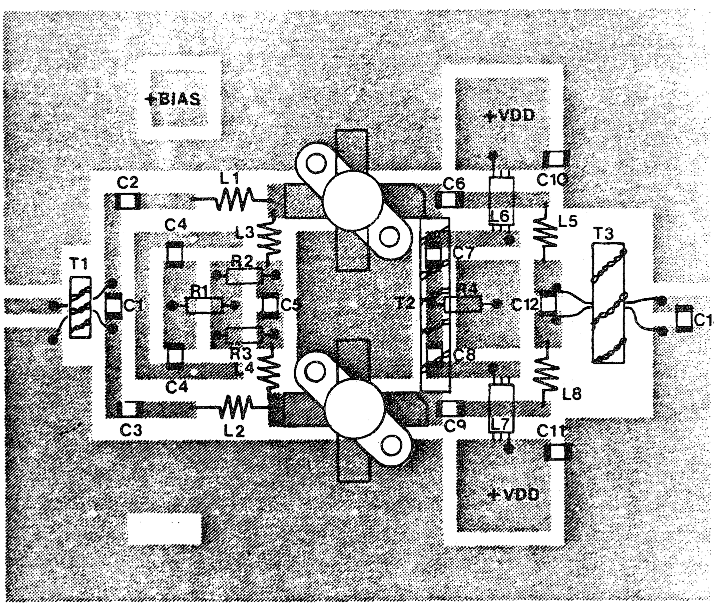

Fig.15 Component lay-out

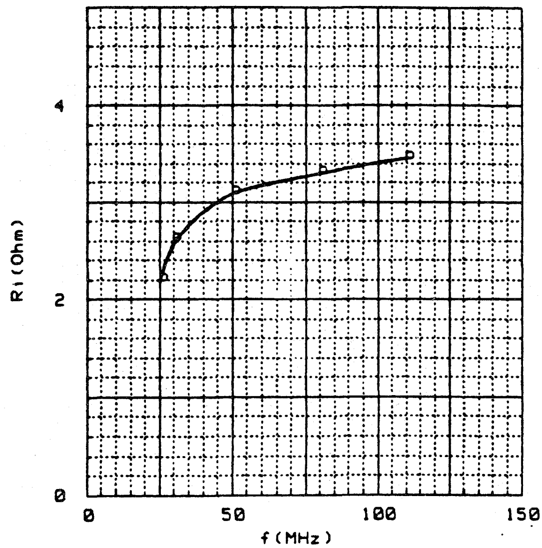

Fig.2 Real part of input impedance

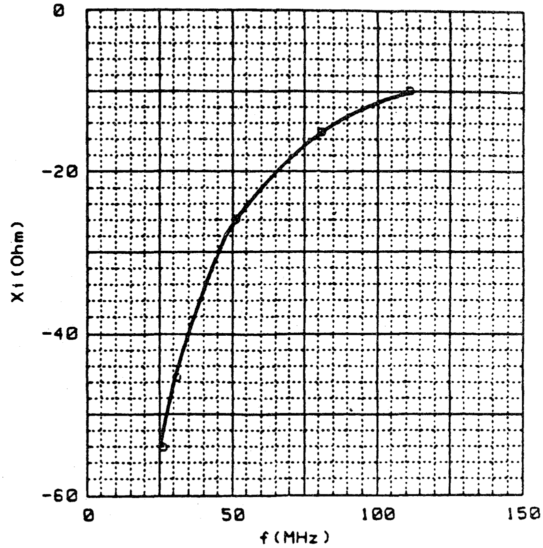

Fig.3 Imaginary part of input impedance

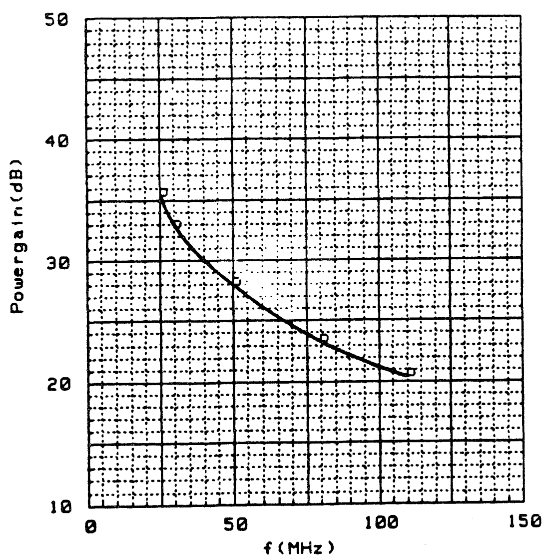

Fig.4 Powergain of transistor

Conditions: Vds = 28 Volt
Idq = 25 mA
Po = 15 Watt
Th = 25 °C

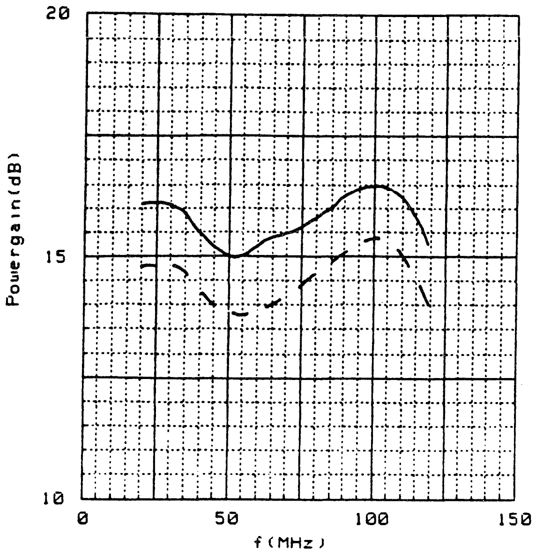

Fig.18 Powergain versus frequency

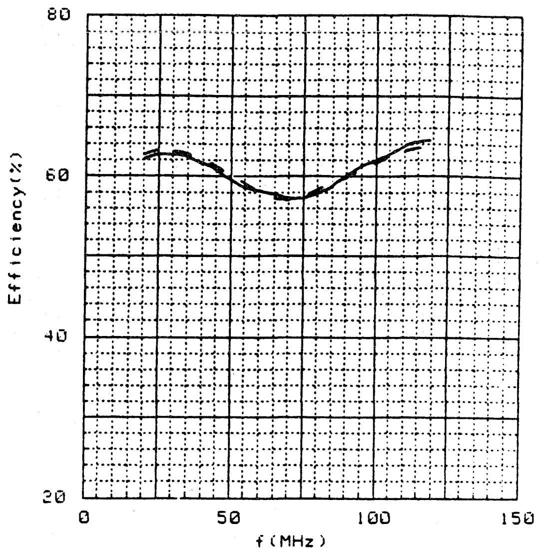

Fig.19 Drain efficiency versus frequency

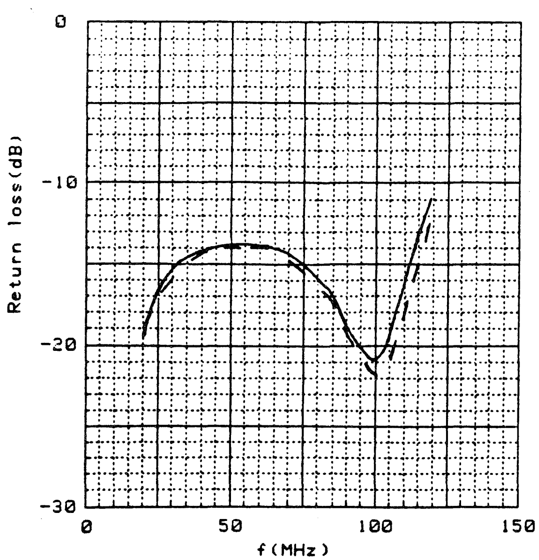

Fig.20 Return loss versus frequency

Conditions: Vds = 28 Volt
Idq = 50 mA
Po = 30 Watt

——— Th = 25 °C
— — — Th = 70 °C

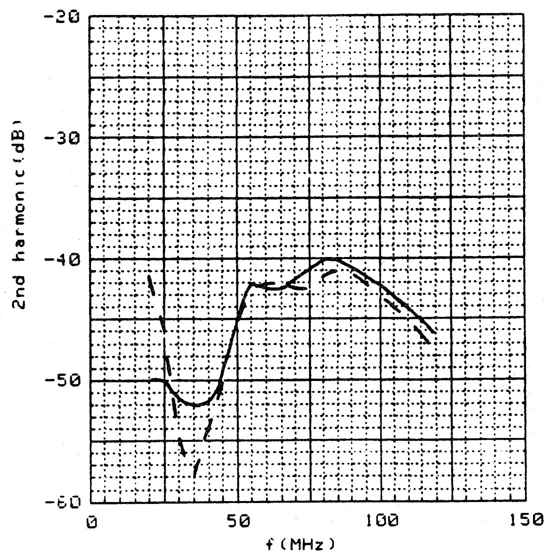

Fig.21 2nd harmonic level versus frequency

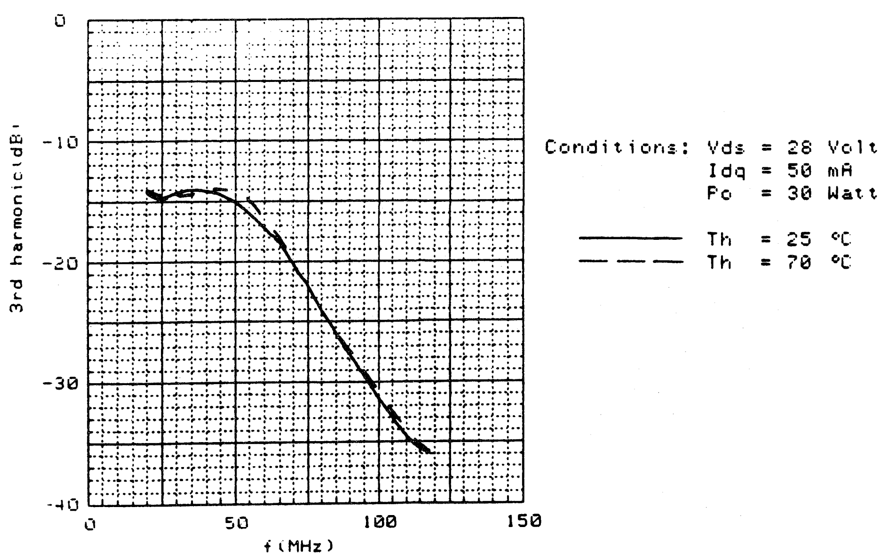

Fig.22 3rd harmonic level versus frequency

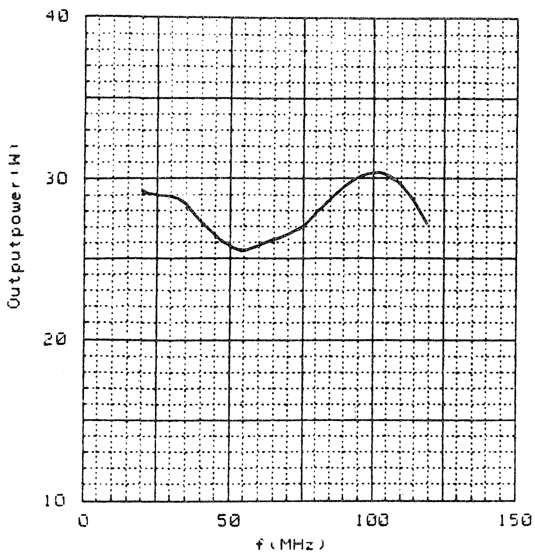

Fig.23 Output power versus frequency

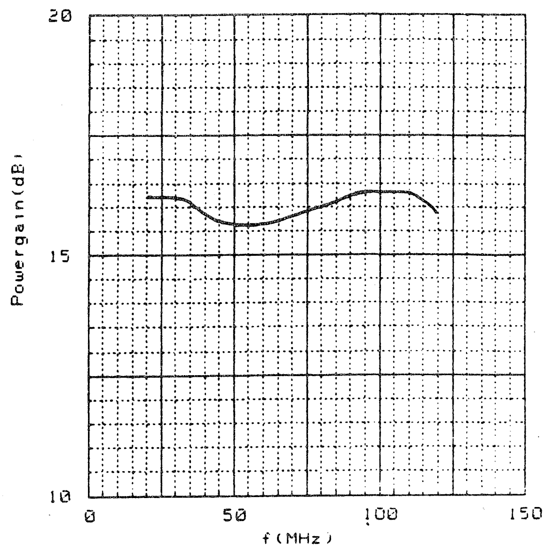

Fig.24 Powergain versus frequency

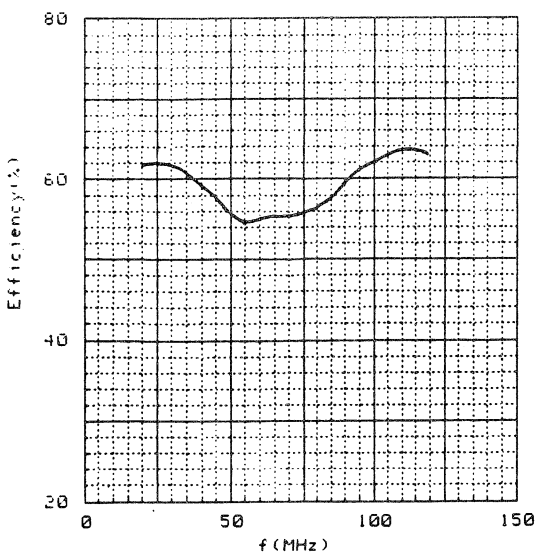

Fig.25 Drain efficiency versus frequency

Conditions: Vds = 28 Volt
Idq = 50 mA
Pin = 700 mW
Th = 25 °C

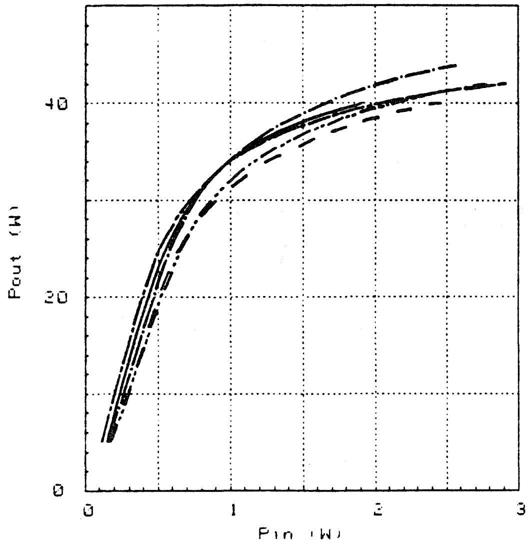

Fig.26 Output power versus input power

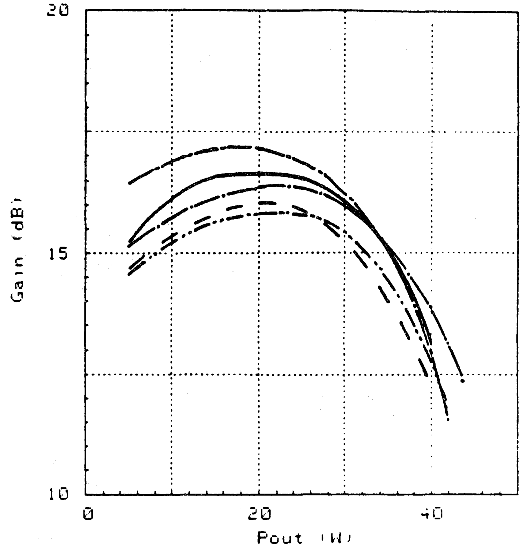

Fig.27 Powergain versus output power

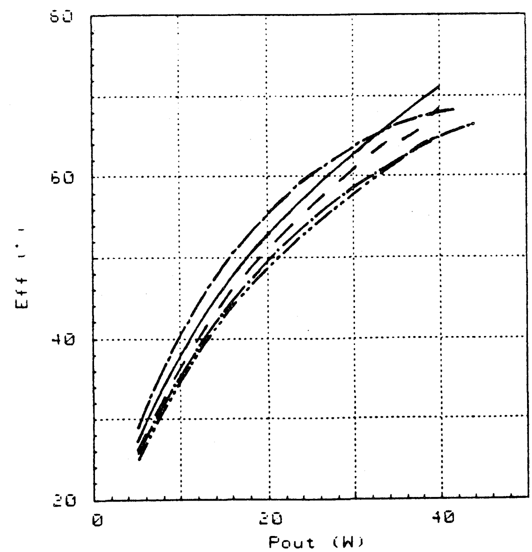

Fig.28 Drain efficiency versus output power

f=25MHz
f=45MHz
f=65MHz
f=85MHz
f=110MHz

Conditions: Vds = 28 Volt
Idq = 50 mA
Th = 25 °C

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APPLICATION

Report no : RNR-1-238-1987-AS / NCO 8702
Author : J.Gajadharsing
Date : 1987-04-16

LINEAR PERFORMANCE AND NOISE FIGURE OF THE WIDEBAND 30WATT PUSH-PULL AMPLIFIER FOR THE FREQUENCY RANGE 25-110MHz WITH TWO MOS TRANSISTORS BLF244.

INTRODUCTION

In application report RNR-1-149-1987-AS a description is given of a wideband power amplifier for the frequency range 25-110MHz. This amplifier is primarily designed for non-linear operation at an output power of 30 Watt. When this amplifier is used in VHF frequency hopping equipment for military communication purposes linearity is required to some extent, d_3 <-26dB.
Also its noise performance is important. To investigate these aspects of the amplifier additional measurements have been carried out.

LINEAR PERFORMANCE

A two-tone intermodulation distorrion test has been performed on this amplifier. The tone separation was 10KHz. Measurements were first carried out under nominal conditions, which are:

Supply voltage V_{dd} = 28 Volt
Quiescent drain current I_{dq} = 50mA/transistor
Heatsink temperature T_{hs} = 25°C

It appeared that the third-order intermodulation distortion product IMD(d3) was the strongest in the output signal. This was <-22.5dB with reference to one of the two tones at the high end of the band. It also appeared that the IMD products increased at lower power levels.
Linearity measurements were repeated at a higher quiescent drain current corresponding with class-AB operation.
At I_{dq}=12.5mA transistor IMD(d3) improved to <-26.5dB.
It also turned out that the IMD products at lower power levels did not increase as in the previous case but stayed nearly constant or decreased.
Page 3 and 4 show IMD (d3) versus the peak envelope output power at four different frequencies. Curves are given for both I_{dq}=50mA and I_{dq}=125mA.
On page 5 powergain and efficiency versus frequency are given based on two-tone measurements.
At I_{dq}=125mA the powergain is approx. 1dB higher than the powergain at I_{dq}=50mA. The drain efficiency is approx.3% lower.

Input return loss and output second harmonic suppression at I_{dq}=125mA were better than -13.5dB resp. -37.5dB.

NOISE PERFORMANCE

Noise measurements, were performed at several frequencies in the band. Laboratory facilities restricted the lowest frequency of measurement to 40MHz. The table below contains the measured noise figure of the amplifier at four frequencies obtained with the noise generator method.
The quiescent drain current was set to 500mA per transistor.

Freq. (MHz)	F (dB)
40	4.8
50	4.3
90	4.4
110	4.3

CONCLUSIONS

This amplifier can only meet the linearity requirement of <-26dB over the whole frequency range 25-110MHz if it is operated in class-AB with I_{dq}=125mA.
When operation is considered only in the band 30-88MHz its IMD is better than -31dB.
The measured noise figure at I_{dq}=500mA is 4-5dB.

J.Gajadharsing

Electronic
components
and materials

PHILIPS

laboratory report

central application laboratory CAB
eindhoven - the netherlands

number : ECO 7804	date : 31.07.1978
project : 6798	pages: A1 ; S2 ; R11

title A 3-STAGE 10W POWER AMPLIFIER FOR THE 470 MHZ BAND, OPERATING FROM 13,5V SUPPLY VOLTAGE

author A. Boekhoudt

ABSTRACT

For application in mobile radio transmitters a 3-stage power amplifier has been designed for the 400-470 MHz frequency band. An output power of 10W is produced at a supply voltage of 13,5V. The following high-gain U.H.F. transistors have been used: BFR95-BLW80-BLW81.
The required drive power is 25mW and the overall efficiency is between 53 and 59%.
The amplifier is stable for load mismatches up to a VSWR of 8.7 (any phase) combined with any input drive up to an output power of 10W.

appr. R.A. Pölzl

Advies Octrooi d.d. 22 aug. 1978	X AV	GV		B		BL
Opgave Mamo d.d. -7 aug. 1978	X AV	X GV	X SP	B		BL
Datum: -1 aug. 1978	Mamo					

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Electronic components and materials

SUMMARY

Figure A shows the detailed circuit diagram. It is a three stage power amplifier for the frequency range of 400 to 470 MHz with the pre-driver and driver transistors operating in class AB and the final transistor in class B. The transistors applied are: BFR95, BLW80 and BLW81 at a supply voltage of 13,5V.

To reduce the variation of output power with supply voltage an automatic control circuit for the supply voltage of the pre-driver has been applied.

In the output network of the amplifier a stripline circuit has been used, the other matching networks are built with conventional inductors and capacitors.

The amplifier is made on PTFE glass-fibre P.C.-board, copper clad on both sides.

Some typical results are: P_i = 25mW, P_o = 10W; gain 26dB.

f (MHz)	I_{tot} (A)	eff (%)
400	1,3	57,0
420	1,25	59,3
440	1,33	55,7
470	1,4	52,9

The amplifier is stable for load mismatches up to a VSWR of 8,7 (any phase) combined with any input drive up to an output power of 10W.

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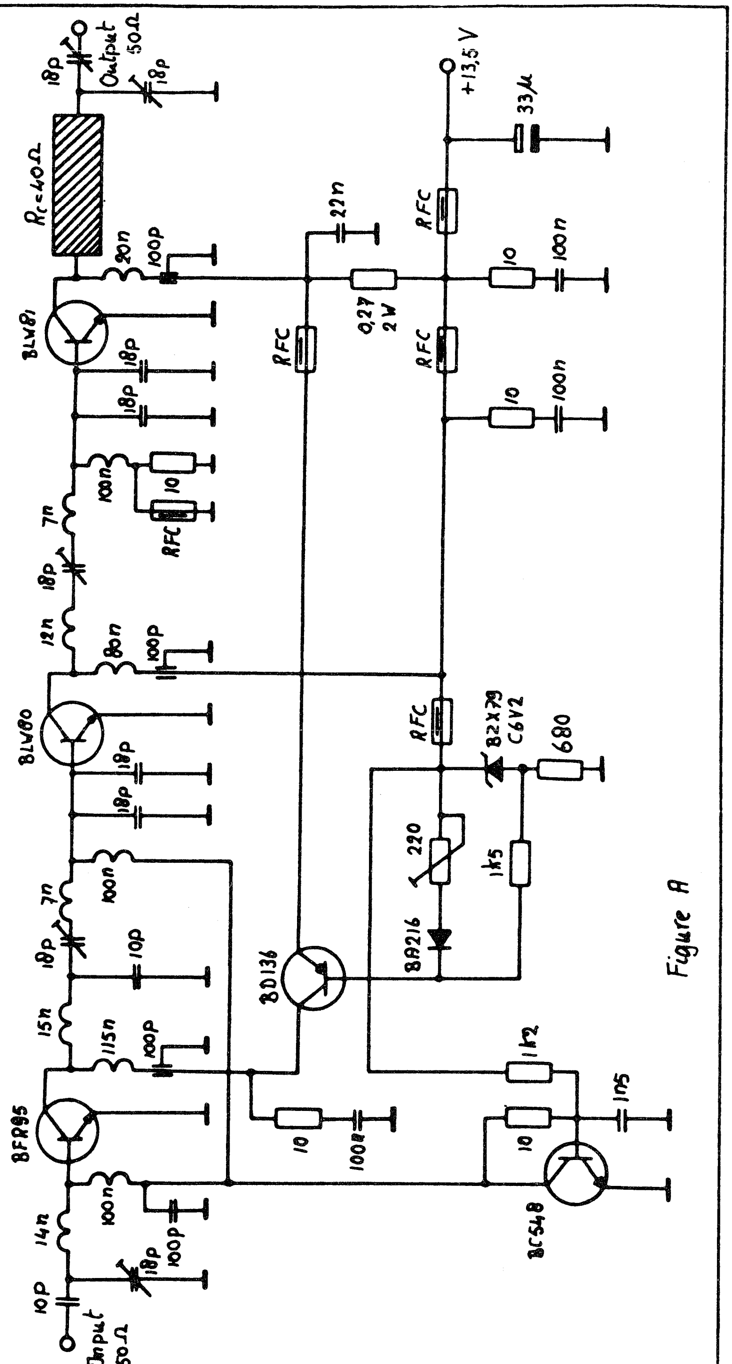

Figure A

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1. INTRODUCTION

The BLW80 and BLW81 are new U.H.F. power transistors to be used in mobile radio transmitters. Together with the BFR95 a three stage amplifier can be built for 10W output power with a drive power of 25mW. The supply voltage is 13,5V and the frequency range is from 400 to 470 MHz.
The BFR95 is derived from the BFR94 and mounted in a TO-39 envelope with the collector connected to the case.
The BLW80 and BLW81 are encapsulated in a ceramic envelope with ¼ inch stud (SOT-122).

2. DESIGN CONSIDERATIONS

The first experiments were done with the transistors BLW79-BLW80-BLW81. In that case the BLW79 had to deliver only 400mW, compared with 2W for which it is specified. To obtain a reasonable stability in this stage the supply voltage had to be reduced to approx. 6V. The voltage drop of 7.5 volt was achieved with a variable series resistor. So it was also possible to control the collector supply voltage.
After this amplifier was realized, we tried to make it cheaper. The BLW79 was replaced by a BFR95. The resistor in the collector supply lines was not omitted because the BFR95 can deliver 1W and so it is possible to control the output power for a constant input power.

A problem with the amplifier was that during variation of the drive power it stopped when the output power became under 3W and it started again for an output power of 5W. This on-off jumping has to be avoided. A solution to this problem was, to adjust the predriver and the driver in class AB.

A 20% reduction of the battery voltage caused a decrease in power of about 50% which is too much.
The solution was to replace the resistor in the collector supply line of the BFR95 by an automatic control circuit (see figure 1).

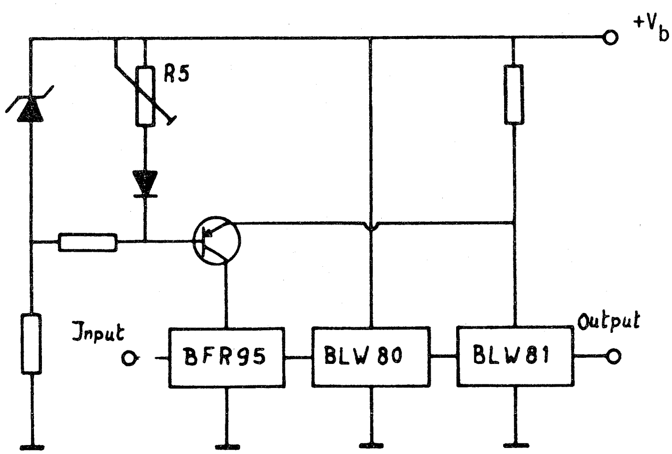

Figure 1

3. CIRCUIT DESCRIPTION

3.1. Output network of the BLW81

The output impedance of the BLW81 at 435MHz is 5,6-j2,5 ohms. Here a stripline is used to transform this impedance to an impedance with a resistive parallel component of 150 ohm, which is matched to 50 ohms with the aid of two trimmers.

3.2. The interstage network between BLW80 and BLW81

The output impedance of the BLW80 is 5,7-j11,2 ohm.
The input impedance of the BLW81 is 1,31+j2,21 ohm.
Both values hold at a frequency of 435 MHz.
Capacitors C_{11} and C_{12} have been dimensioned such that their total reactance is twice the equivalent parallel input reactance of the transistor. The rest of the matching has been made in two steps with an intermediate impedance of 200 ohms at the interconnection point of L_8 and C_{10}. The capacitance from this point to ground is formed by the track on the p.c. board.

3.3. The interstage network between BFR95 and BLW80

The output impedance of the BFR95 is 26-j41 ohm and the input impedance of the BLW80 is 2,1+j2,3 ohm.
The network is calculated in the same way as the interstage network of the BLW80-BLW81.

3.4. The input network of the BFR95

The input impedance is 11,35-j6,2 ohm. The impedance transformation to 50 ohms has been made in the conventional way.

4. CONSTRUCTIONAL DETAILS

The printed circuit board is 1,5 mm (1/16 inch) PTFE-fibre glass, copper clad on both sides. The emitters of the BLW80 and BLW81 are connected to earth by means of copper strip between the upper and lower side of the p.c. board. The BFR95 is mounted on the upper side of the p.c. board while here rivets are used to make contact between both sides.

5. MISMATCH TESTS AND SPURIOUS GENERATION

An important requirement is that the amplifier must be stable and has to survive mismatch tests applied at the output.

The stability tests have been done according to the following specifications:

. measuring frequencies: 400-420-440-470 MHz
. supply voltage V_b = 13,5V
. output power P_o = 10W
. drive level P_i = 25mW
. VSWR (output) 1:8.7 (0-360^o)
. heatsink temperature approx. 20^{o}C

Tests have been made with the set-up shown in figure 2 in which the spectrum analyzer HP8558B plays an important part.
A Rohde and Schwartz signal generator type SMLU has been used.
The frequency counter is a Philips type PM6615.
The forward and reflected power at the input is measured by a Rohde and Schwartz watt-meter and matching indicator, type NAU.
The output power is attenuated 30dB by a Bird attenuator.
The power is measured by a Hewlett-Packard power meter type 435A.
The mismatch tests were done with a normal 1dB power attenuator and a reactance unit.
The currents were measured with a Hewlett-Packard clip-on meter model 428 B.

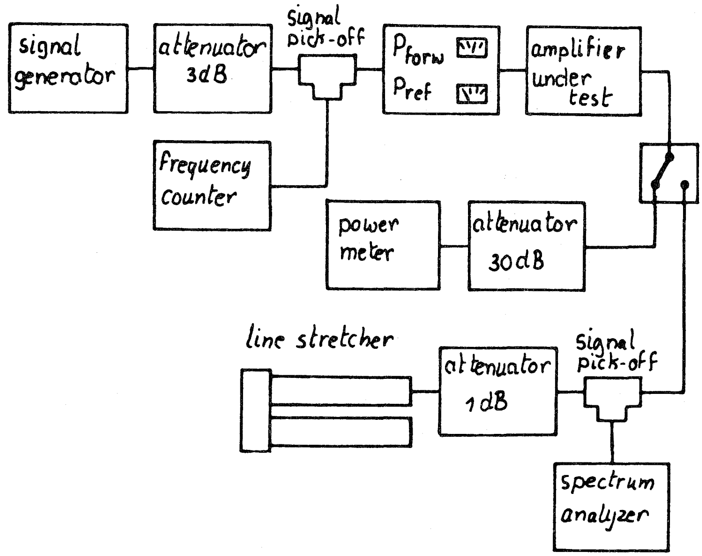

Figure 2. Test set-up

6. MEASUREMENTS

At every test frequency, the interstage and output trimmers were tuned for maximum output power and the input trimmer for minimum reflection.

At the same time the output power was adjusted at 10W by means of R_5.

The results are:

	P_i = 25mW	P_o = 10W Gain = 26dB		
f (MHz)	I_{T1} (mA)	I_{T2} (mA)	I_{TOT} (A)	n (%)
400	19,7	320	1,3	57,0
420	23	266	1,25	59,3
440	31	318	1,33	55,7
470	50	269	1,4	52,9

For an output power drop of 0.5W the instantenous bandwidth is 1.7% and for an output power drop of 1W it becomes 2.5%.
A 20% reduction of the battery voltage causes a maximum output power decrease of 36%.
The amplifier is stable for load mismatches up to a VSWR of 8,7 (any phase) combined with any input drive up to 25 mW.

7. REFERENCES

Three-stage 15W Power Amplifier for the 470 MHz Communication Band, Electronic Applications Bulletin, volume 30, number 4, pp. 121-130, by M.J. Köppen.

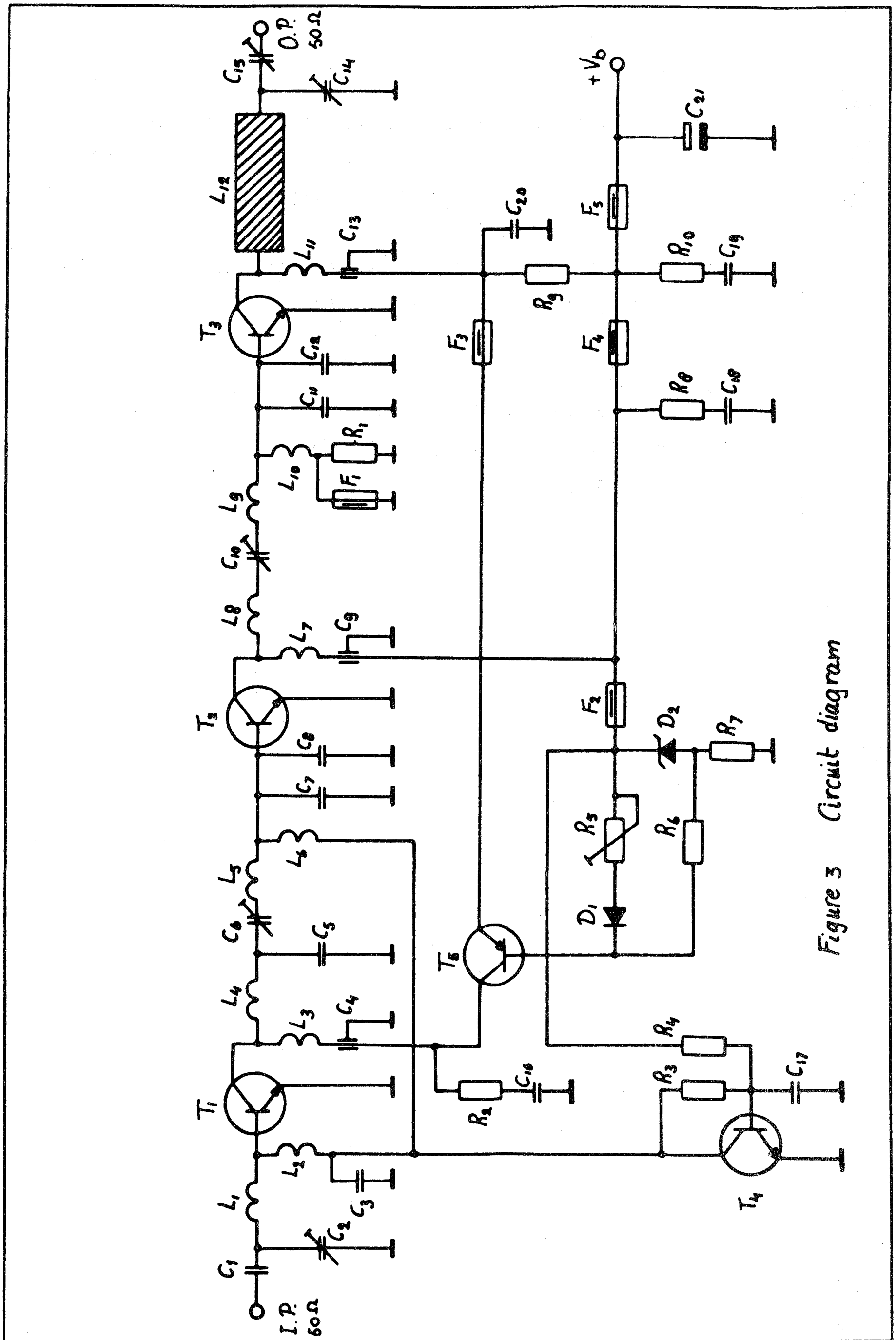

Figure 3 Circuit diagram

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8. PARTS LIST

R_1, R_2, R_3, R_8, R_{10} = 10 ohm, carbon, ± 5%, CR37 style
R_4 = 1,2k ohm, carbon, ± 5%, CR25 style
R_5 = 220 ohm, carbon potmeter
R_6 = 1,5k ohm, carbon, ± 5%, CR25 style
R_7 = 680 ohm, carbon, ± 5%, CR25 style
R_8 = 0,27 ohm, wire wound, ± 10%, 2W

C_1, C_5 = 10pF, 500V ceramic
C_2, C_{14} = 2-18pF, film dielectric trimmer (cat.nr. 2222 809 09003)
C_3 = 100pF, 500V ceramic
C_4, C_9, C_{13} = 100pF, ceramic feed-through capacitors
C_6, C_{10}, C_{15} = 2-18pF, film dielectric trimmer
(cat.nr. 2222 809 05003)
C_7, C_8, C_{11}, C_{12} = 18pF, chip capacitor (cat.nr. 2222 851 13189)
C_{16}, C_{18}, C_{19} = 100nF, polyester, ± 10%
C_{17} = 1,5nF, 500V ceramic
C_{20} = 22nF, 63V ceramic
C_{21} = 33 µF, aluminium electrolytic, 16V

F_1, F_2, F_3, F_4, F_5 = ferroxcube bead with 3 turns of 0,6 mm CuEm wire, cat.nr. of bead 4313 020 15172.

L_1 = 14nH; 0,5 turns, D_{int} = 4,5 mm, d = 1,0 mm Cu wire, leads 2x5mm.
L_2, L_6, L_{10} = 6 turns, D_{int} = 4 mm, d = 0,6 mm CuEm wire, closely wound, leads 2x5mm.
L_3 = 115nH; 5 turns, D_{int} = 5 mm, d = 0,6 mm CuEm wire, closely wound, leads 2x5mm.
L_4 = 15nH; 0,5 turn, D_{int} = 5 mm, d = 1,0 mm Cu wire, leads 2x5mm.
L_5 = 7nH, printed on p.c. board

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L_7 = 80nH; 6 turns, D_{int} = 3,5 mm, d = 0,6 mm CuEm wire, closely wound, leads 2x5mm.

L_8 = 12nH; 0,5 turn , D_{int} = 3.5 mm, d = 1,0 mm Cu wire, leads 2x5mm.

L_9 = 7nH, printed on p.c. board

L_{11} = 20nH; 2 turns, D_{int} = 3 mm, d = 1,1 mm CuEm wire , closely wound, leads 2x5mm.

L_{12} = stripline 58,8 mm x 6 mm ($R_c \simeq$ 38 ohms)

T_1 = BFR95
T_2 = BLW80
T_3 = BLW81
T_4 = BC548
T_5 = BD136

D_1 = BA216
D_2 = BZX79-C6V2

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Electronic components and materials

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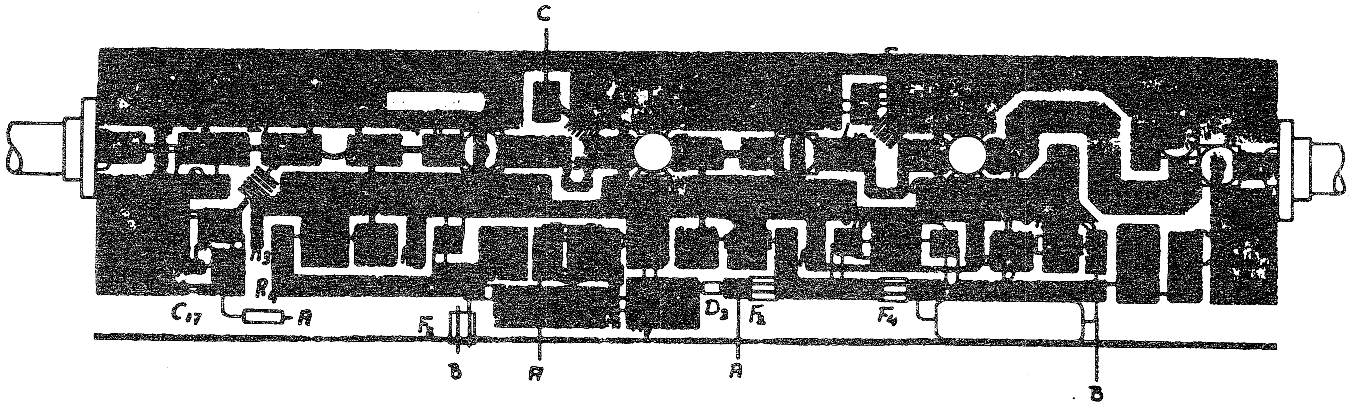

Figure 4 components lay-out

To be sure of good contact between upper and lower side, some rivets are used for the trimmers C_2, C_{14} and the emitter of T_1. Copper foil is used for the emitters of T_2, T_3.

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Electronic components and materials

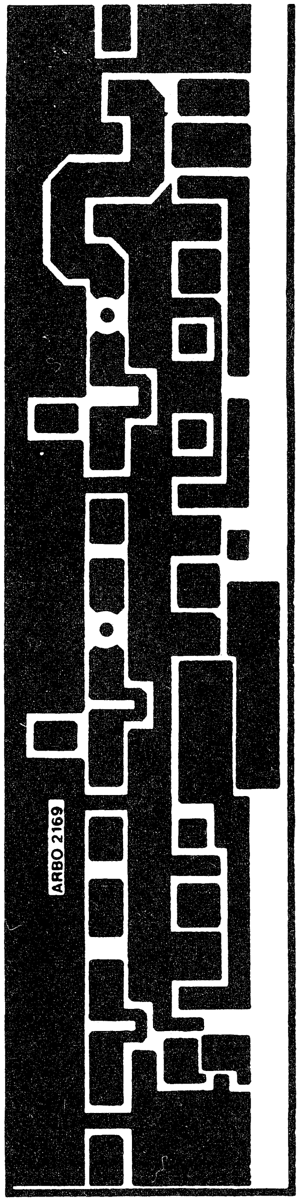

Figure 5 print lay-out

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Electronic
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laboratory report

number : EC07805 date : 03-08-1978

title : A 30W BOOSTER AMPLIFIER FOR THE 470MHz BAND, OPERATING FROM 13,5V SUPPLY VOLTAGE.

author : A.Boekhoudt

PHILIPS Electronic components and materials

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laboratory report

central application laboratory CAB
eindhoven - the netherlands

number :	ECO 7805	date :	03.08.1978
project :	6798	pages:	A1 ; S1 ; R6

title A 30W BOOSTER AMPLIFIER FOR THE 470MHz BAND, OPERATING FROM 13,5V SUPPLY VOLTAGE

author A. Boekhoudt

ABSTRACT

The 30W amplifier described in this report is ment as a booster for a 10W mobile radio transmitter in the 400-470MHz frequency band. It has been tested in combination with an already existing 3-stage amplifier for this band. The power transistor applied is the BLW82 at a supply voltage of 13,5V.
The combined amplifiers have a power gain of nearly 31dB and they are stable with load mismatches up to a VSWR of 8,7 (any phase) combined with any input drive up to an output power of 30W.

R.A. Pölzl

Advies Octrooi d.d. 22 aug. 1978	X AV	GV		B.......		BL
Opgave Mamo d.d. 14 aug. 1978	X AV	X GV	X SP	B.......		BL
Datum: -9 aug. 1978	Mamo					

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SUMMARY

The UHF power amplifier (400-470MHz) described in this report is ment as a booster for the three-stage UHF amplifier described in report ECO 7804. The output power of the booster is 30W. It operates in class B and the transistor used is the BLW82 with a supply voltage of 13,5V.
In the input as well as in the output network striplines have been used. The amplifier is printed on a double copper clad PTFE glass-fibre board.

Some typical results of the four stage amplifier are: P_i = 25mW; P_o = 30W; Gain = 30,8dB.

f(MHz)	I_1(A)	I_{tot}(A)	eff_1(%)	overall eff(%)
400	4,3	5,15	52,3	43,1
420	3,3	4,15	67,3	53,5
440	3,7	4,6	60,1	48,3
470	3,4	4,5	65,4	49,4

I_1 and eff_1 are respectively the collector current and the efficiency of the BLW82.

The amplifier is stable for load mismatches up to a VSWR of 8,7 (any phase) combined with any input drive up to an output power of 30W.

1. Introduction

In report ECO 7804 a description has been given of a three-stage UHF amplifier for 10 Watts output and a drive power of 25mW, using the transistors BFR95, BLW80 and BLW81. In this report application information will be given on a booster amplifier for the above mentioned one. In this booster the BLW82 has been applied. It produces an output of 30W over the frequency range of 400 to 470MHz. The supply voltage is 13,5V. The BLW82 is encapsulated in a 6 leads ½ inch flange envelope with a ceramic cap. It contains a built-in matching section at the input to raise the real part of the input impedance, by which the Q-factor of this impedance is reduced.

2. Circuit description

The input impedance of the BLW82 at 435MHz is 1,3+j2,9 ohms and the output impedance 2,6+j1,7 ohms. Both impedances are matched to 50 ohms in two steps. In both cases up-transformation has been achieved by means of a stripline, which has been dimensioned such that at the end of the line the resistive parallel component of the impedance is between 100 and 150 ohms. These impedances are matched to 50 ohms by series and parallel capacitors, see figure 1.

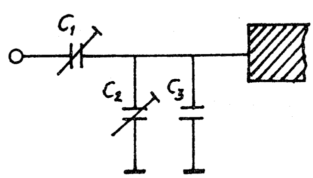

figure 1

3. Construction

The circuit used as booster was originally a test circuit for 470MHz, but to make it also suitable for 400MHz, some additional capacitance is required in parallel with C_2-C_3 and C_6-C_7 (see figure 3). A better solution for the capacitors C_2-C_3-C_9 and C_6-C_7-C_{10} is to use a trimmer with a maximum capacitance of 18pF in parallel with ceramic capacitor of 3,3pF.
During mismatch tests it appeared that the amplifier produced some spurious oscillations. The remedy was to put in series with the base choke, the parallel connection of a 10 ohm resistor and a ferroxcube choke.
A copper strip under the emitters is used to get a good contact between upper and lower side of the PC board.

4. Mismatch tests and spurious generation

The amplifier has to be stable for an output VSWR of 8,7 (0-360^o). The stability tests have been done according to the following specifications:

. measuring frequencies: 400-420-440-470MHz.
. supply voltage V_b = +13,5V.
. output power P_o = 30W.
. drive level P_i = 25mW (input to BFR95)
. VSWR (output) 1:8,7 (0-360^o)
. heatsink temperature approx. 20^{o}C.

Tests have been made with the set-up shown in figure 2.

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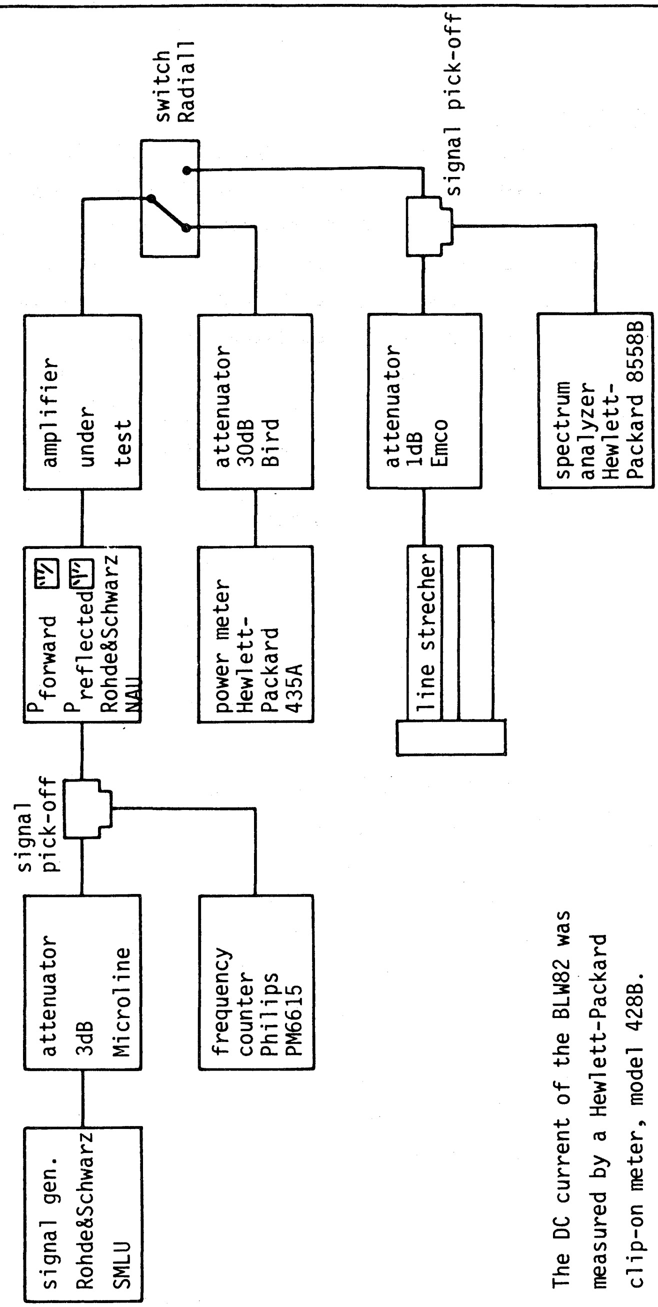

The DC current of the BLW82 was measured by a Hewlett-Packard clip-on meter, model 428B.

figure 2

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5. Measurements

At every test frequency the three-stage amplifier was tuned for maximum output power at 50 ohms input and ouput impedance.
In this situation the booster was connected and also tuned to maximum output power with C_1-C_2 and C_7-C_8. At the same time the output power was adjusted at 30W by means of variable resistor R5 of the 3-stage amplifier (see reference). The results are:

P_i = 25mW	P_o = 30W	Gain = 30,8dB	I_1 = DC current consumption of BLW82 eff_1 = efficiency of booster	
f(MHz)	I_1(A)	I_{tot}(A)	eff_1(%)	Overall eff(%)
400	4,3	5,15	52,3	43,1
420	3,3	4,15	67,3	53,5
440	3,7	4,6	60,1	48,3
470	3,4	4,5	69,4	49,4

The instantaneous bandwidth of the amplifier is about 1% for an output power drop of 5%.
For a reduction of 20% of the supply voltage, the output power will have a maximum drop of 45%.
The amplifier is stable for load mismatches up to a VSWR of 8,7 (any phase) combined with any input drive up to 25mW.

Reference:

Report no. ECO 7804 by A. Boekhoudt.
"A 3-stage 10W power amplifier for the 470MHz band, operating from 13,5V supply voltage".

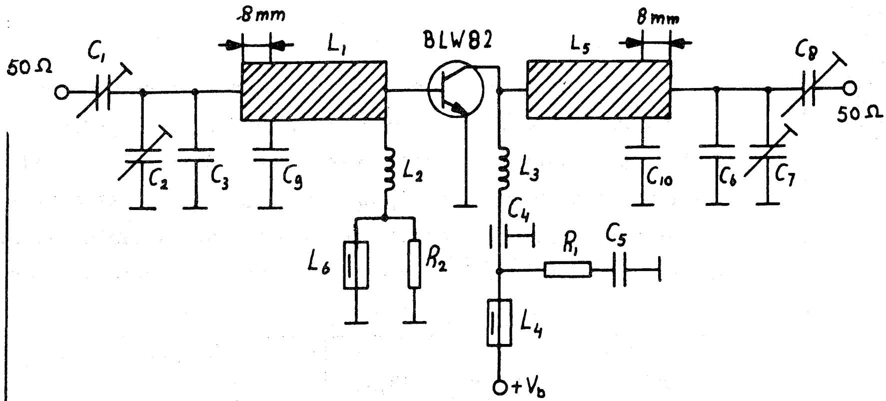

figure 3

Parts list

$R_1 = R_2$ = 10 ohm, carbon resistor.

$C_1 = C_2 = C_7 = C_8$ = 2 to 9pF, film dielectric trimmer (cat.no. 2222 809 09002).

$C_3 = C_6$ = 3,9pF, ceramic capacitor (500V).

C_4 = 100pF, feed through capacitor (cat.no. 2222 700 04101).

C_5 = 100nF, polyester capacitor.

C_9 = 4,7pF, ceramic capacitor (500V).

C_{10} = 6,8pF, ceramic capacitor (500V).

L_1 = stripline (24,0 x 6,7mm), $R_c \simeq$ 35 ohms.

L_2 = 10 turns closely wound enamelled Cu wire (0,4mm); int. dia 4mm.

L_3 = 12,6nH; 2,5 turns enamelled Cu wire (0,7mm); int. dia 4mm; length 3mm; leads 2 x 5mm.

L_4 = Ferroxcube choke (cat.no. 4312 020 36640).

L_5 = Stripline (28,4 x 6,7mm), $R_c \simeq$ 35 ohms.

L_6 = Ferroxcube bead with 2 turns of 0,6mm enamelled Cu wire (cat.no. 4313 020 15170).

L_1 and L_5 are striplines on a double copper clad printed circuit board with PTFE fibre-glass dielectric (ε_r = 2,74); thickness 1/16".

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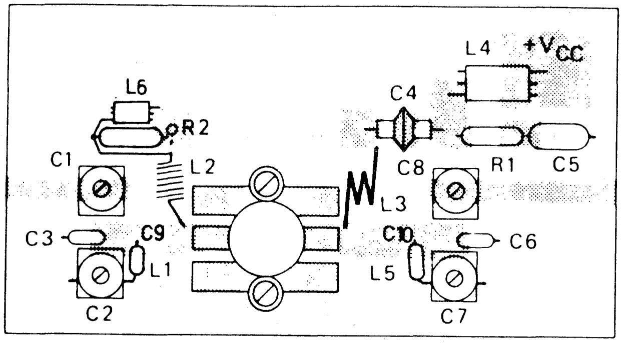

7Z78204

Figure 4 component lay-out

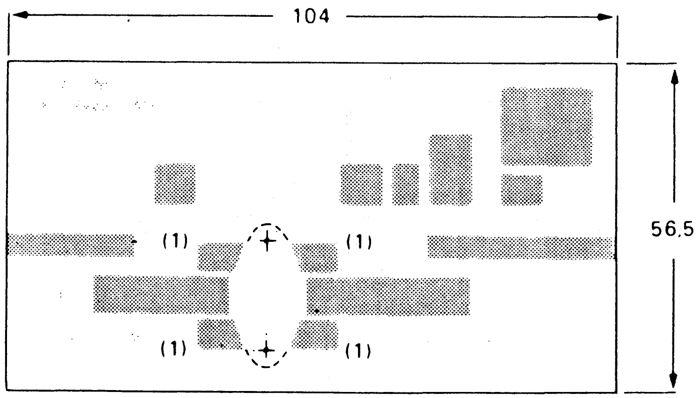

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Figure 5 print lay-out

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laboratory report

central application laboratory CAB
eindhoven - the netherlands

number : ECO 8102 date : 1981-11-19

project : 8092 pages: A1 ; R3 ;

title THE BFR90A, BFR91A AND BFR96S AS DRIVER AMPLIFIERS IN MOBILE AND PORTABLE RADIO TRANSMITTERS

author A.H. Hilbers

ABSTRACT

In this report input and load impedance and power gain have been given for the above mentioned transistors as class-B driver amplifiers in small V.H.F. and U.H.F. transmitters.
These data have been given under the following conditions which must be considered as maxima for safe operation:

Type	V_{CE} (V)	P_o (mW)
BFR90A	10	100
BFR91A	7,5	160
BFR96S	10	500

Appr. R.A. Pölzl

Advies Octrooi d.d. 30-11-81	AV X	GV		B......		BL
Opgave Mamo d.d. 30 11-81	AV	GV X	SP	B......		BL
Datum: 19 NOV. 1981	Mamo					

1. INTRODUCTION

From time to time we receive questions about the application possibilities of our wideband transistors as class-B driver and pre-driver stages in the transmitters of mobile and portable radios.

For this purpose a theoretical investigation has been done concerning:

a. maximum supply voltage
b. maximum output power
c. optimum load impedance, input impedance and power gain in the frequency range of 25 to 500 MHz.

As the f_T of these transistors is rather high in comparison with the operating frequencies it is necessary to apply a damping resistor between the base and emitter terminals of the transistor to achieve sufficient stability.

On the following pages the above mentioned information will be given for the BFR90A, BFR91A and BFR96S.

It must be noted that in the figures for power gain and input impedance the influence of the base-emitter damping resistor has already been taken up.

2. BFR90A

V_{CE} = 10V (max)
P_0 = 100mW (max)
R_{BE} = 100 Ohm (damping resistor)

f MHz	G dB	Inp. Imp. Ohm	Load Imp. Ohm
25.0	13.41	98.39 - j 9.70	467.95 + j 215.05
50.0	13.57	99.25 - j 19.59	351.82 + j 272.96
100.0	13.92	95.95 - j 39.86	223.39 + j 274.39
150.0	14.29	84.42 - j 56.97	159.73 + j 251.94
200.0	14.63	67.94 - j 66.55	123.33 + j 230.65
300.0	15.08	38.81 - j 65.26	85.14 + j 199.28
400.0	15.12	23.36 - j 54.18	66.41 + j 178.16
500.0	14.68	16.43 - j 43.95	55.20 + j 163.43

Table 1

3. BFR91A

V_{CE} = 7.5V (max)
P_0 = 160mW (max)
R_{BE} = 82 Ohm (damping resistor)

f MHz	G dB	Inp. Imp. Ohm	Load Imp. Ohm
25.0	14.31	79.11 - j 9.57	176.95 + j 59.08
50.0	14.40	77.94 - j 19.03	146.70 + j 85.58
100.0	14.66	69.58 - j 35.45	105.68 + j 96.81
150.0	14.84	55.80 - j 44.56	84.23 + j 95.76
200.0	14.93	42.27 - j 46.07	71.96 + j 92.75
300.0	14.72	24.99 - j 38.94	59.53 + j 88.16
400.0	14.05	17.35 - j 30.19	53.30 + j 84.57
500.0	13.10	13.91 - j 23.32	49.09 + j 82.34

Table 2.

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4. BFR96S

V_{CE} = 10V (max)
P_0 = 500mW (max)
R_{BE} = 47 Ohm (damping resistor)

f	*	G	*	Inp. Imp.	*	Load Imp.
MHz	*	dB	*	Ohm	*	Ohm
25.0	*	16.66	*	44.40 - j 8.50	*	97.66 + j 43.65
50.0	*	16.84	*	41.77 - j 16.45	*	77.15 + j 55.34
100.0	*	17.17	*	29.95 - j 25.26	*	57.65 + j 57.45
150.0	*	17.23	*	19.09 - j 24.42	*	50.76 + j 56.80
200.0	*	16.99	*	12.88 - j 20.34	*	47.96 + j 56.01
300.0	*	15.77	*	8.02 - j 12.87	*	45.94 + j 54.92
400.0	*	14.08	*	6.65 - j 7.79	*	44.72 + j 54.43
500.0	*	12.32	*	6.31 - j 4.13	*	43.31 + j 54.10

Table 3.

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N.V. PHILIPS SEMICONDUCTORS NIJMEGEN – THE NETHERLANDS

PRODUCT GROUP SPECIALTIES AND DIODES

Report no.: NCO-8402
Author : A.H.Hilbers
Date : 1984-04-09

RNR-1-167-1984-AS
APPLICATION

THE BFG 90A,BFG 91A AND BFG 96 AS DRIVER AMPLIFIERS IN MOBILE AND PORTABLE RADIO TRANSMITTERS

ABSTRACT

For the above mentioned transistor types power gain, input and load impedance have been given for application as class-B driver stages in the frequency range 400 to 1000 MHz.

This information is given under the following conditions which must be considered as maxima for safe operation:

TYPE	V_{CE} (V)	P_o (mW)
BFG 90 A	10	100
BFG 91 A	7.5	160
BFG 96	10	500

INTRODUCTION

In 1981 we have given design information for use of the BFR 90A, BFR 91A and BFR 96S in driver stages of mobile and portable transmitters up to a frequency of 500MHz.
Last year we have introduced so-called cross-pack versions of these devices: BFG 90A, BFG 91A and BFG 96.
They have 2 emitter connections instead of one by which they are suitable for use at higher frequencies, so the 800-960MHz communications band can now be included.

The maximum supply voltage and output power of the BFG types are the same as those for the BFR types. A remark has to be made about the BFG 96:
In the development sample data a figure is mentioned for the maximum power dissipation of 500mW being equal to that of the BFR 96.

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In the meantime thermal resistance measurements have been done, showing that this figure can be increased to 700mW like that of the BFR 96 S.

Based on this figure gain and impedance information will be given for an output power of 500mW. This can only be done in situations where no severe mismatch is expected like driver stages. In other situations we recommend the more rugged BLU 98 which can also be used at higher supply voltages up to 15.5V.

The f_T of all 3 types is in the range of 5 to 6GHz.
This makes it necessary to apply some damping at the input of the transistors in the form of a resistor between base and emitter terminals. In this way the stability is improved especially at the lower frequencies.

On the next pages information will be given on power gain, input and load impedance of the afore mentioned transistor types in class-B operation, i.e. V_{BE}=0 (D.C.).

It must be noted that in the figures for power gain and input impedance the influence of the base-emitter damping resistor has already been taken up.

BFG 90 A

V_{CE}=10V (max.)

P_{o} =100mW (max.)

R_{BE}=82 Ohm (damping resistor)

f (MHz)	G_p (dB)	Inp.imp. (Ohm)	Load imp. (Ohm)
400	14.7	25.2 - j 51.8	62.8 + j 175
500	14.9	16.0 - j 43.5	49.7 + j 156
600	14.8	11.4 - j 35.9	41.4 + j 143
700	14.3	9.1 - j 29.8	35.6 + j 133
800	13.6	8.0 - j 24.9	31.4 + j 124
900	12.7	7.4 - j 21.0	27.8 + j 117
1000	11.8	7.3 - j 17.7	25.0 + j 111

../4

BFG 91 A

V_{CE}=7.5V (max.)
P_o =160mW (max.)
R_{BE}=82 Ohm (damping resistor)

f (MHz)	G_p (dB)	Inp.imp. (Ohm)	Load imp. (Ohm)
400	15.8	12.9 - j 33.5	40.5 + j 77.9
500	15.4	9.2 - j 25.6	36.1 + j 74.4
600	14.6	7.5 - j 19.6	33.0 + j 71.6
700	13.6	6.9 - j 14.9	30.5 + j 69.2
800	12.5	6.7 - j 11.1	28.5 + j 67.1
900	11.4	6.7 - j 7.9	26.6 + j 65.2
1000	10.3	6.9 - j 5.2	25. + j 63.3

BFG 96

V_{CE}=10V (max.)
P_o=500mW (max.)
R_{BE}=39 Ohm (damping resistor)

f (MHz)	G_p (dB)	Inp.imp. (Ohm)	Load imp. (Ohm)
400	17.1	3.7 - j 8.8	32.1 + j 50.3
500	15.7	3.3 - j 4.9	30.6 + j 49.4
600	13.8	3.4 - j 1.9	29.3 + j 48.4
700	11.9	3.8 + j 0.5	27.9 + j 47.6
800	10.1	4.3 + j 2.6	26.4 + j 46.6
900	8.4	4.9 + j 4.3	25.0 + j 45.6
1000	6.9	5.5 + j 5.9	23.4 + j 44.6

N.V. PHILIPS SEMICONDUCTORS NIJMEGEN – THE NETHERLANDS

PRODUCT GROUP SPECIALTIES AND DIODES

Report no.: NCO 8403
RNR-1-284-1984-AS
APPLICATION
Author : A.H.Hilbers
Date : 1984-06-28

A 3 STAGE LINE-UP FOR THE 800MHz BAND WITH 8W OUTPUT POWER AT 12.5V

ABSTRACT

A 3 stage power amplifier for mobile radio transmitters in the frequency band from 820 to 870 MHz has been described.
The output power is 8W at a supply voltage of 12.5V.
The amplifier uses the following transistor types:
BLU 98, BLV 91 and BLV 93.
The required drive power is 55mW max. and the overall efficiency is better than 39%.

1. INTRODUCTION

In our Semiconductor Handbook part S6, 1984 we give a number of line-ups for mobile radio transmitters in the 800-900MHz band.
They are intended for a supply voltage of 12.5V.
The first one of these line-ups with an output power of 8W is theoretically investigated for the frequency band of 820-870MHz and is described in this report.
The individual stages have been designed on basis of 50 ohms input and load impedances. In the schematic diagram of Fig.1 the intermediate 50 ohm points have been indicated by the characters A and B.
In this way the stages can be tested and aligned individually.

2. THE OUTPUT STAGE

This stage has been built around the BLV 93 which is a JO common-emitter transistor in a 6 leads flange package with rectangular ceramic cap. (SOT 171). The transistor is specified at 900MHz in class-B for 8W output power at 12.5V:
Power gain: 6.5dB min., 7.3dB typ.
Drive power: 1.8W max., 1.5W typ.

../2

Efficiency: 50% min., 58% typ.
Collector current: 1.28A max., 1.1A typ.

The input and optimum load impedance in this frequency band are given in the table below:

f(MHz)	Z_i (Ohms)	Z_L (Ohms)
820	2.97 + j 2.37	6.69 + j 0.58
845	3.12 + j 2.37	6.52 + j 0.50
870	3.29 + j 2.33	6.33 + j 0.40

The circuit has been designed for 1/32" PTFE fibre-glass with a relative dielectric constant of 2.74.(See parts list).

For the alignment of the output network it can be useful to replace the transistor by a dummy load which is connected between the collector and emitter terminals.
This load must consist of the parallel connection of a 9.1 Ohm chip resistor and a 12pF chip capacitor. The output circuit must then be aligned such that the VSWR measured at the 50 Ohms output terminal is minimum.

To align the input circuit the BLV 93 can be temporarily adjusted in class-A at I_C=1.1A.
Because of the built-in emitter resistance the transistor can withstand this for a longer time provided it is mounted on an adequate heatsink.
The class-A biasing can be achieved by disconnecting L14 at the ground side and by applying a small positive voltage at this point, which must of course be well decoupled.
The input VSWR can now be measured at point B on a small signal basis and the circuit must be aligned to minimize this VSWR.
This method is intended as a coarse alignment. Final alignment must be done in class-B at the actual power level.
In general it can be said that both value and position of C14 and C17 are rather critical, C17 controls the efficiency over the band and C14 the input VSWR.

../3

3. THE DRIVER STAGE

This stage uses the BLV 91, a common emitter transistor in a 4 leads stud package with ceramic cap (SOT 172). The transistor is used in class-B with a maximum output power of 1.8W at 12.5V.

Under these conditions and at 870 MHz we can expect:

Power gain: 7.0 dB min, 8,3 dB typ

Drive power: 0.36W max, 0.27W typ

Efficiency: 50% min, 60% typ

Collector current: 0,29A max, 0.24A typ

The input and optimum load impedance are given in the table below.

f(MHz)	Z_i (Ohms)	Z_L (Ohms)
820	4.03 + j 2.44	18.2 + j 15.7
845	4.04 + j 2.78	18.4 + j 15.8
870	4.03 + j 3.12	18.5 + j 15.9

The alignment of this stage can be done in the same way as that of the output stage.

For the dummy load a parallel connection of 39 ohms and 4.7 pF is recommended and for temporary class-A adjustment a collector current of 0.24A can be chosen.

A 50 ohm source must be connected to point A via a blocking capacitor (27pF, chip).

4. THE PRE-DRIVER STAGE

In this stage a BLU 98 is used. This is a common emitter transistor is a 4 leads plastic pill package (SOT 103). It is used in class-B with a maximum output power of 360mW at 12.5V.

Under these conditions and at 870 MHz we can expect:

Power gain: 8.2 dB min, 9.7 dB typ

Drive power: 55 mW max, 39 mW typ

Efficiency: 50% min, 65% typ

Collector current: 58 mA max, 44 mA typ

The input and optimum load impedance are given in the table below.

f(MHz)	Z_i (Ohms)	Z_L (Ohms)
820	2.78 + j 1.11	10.7 + j 44.7
845	2.94 + j 1.67	10.9 + j 44.8
870	3.08 + j 2.21	10.9 + j 45.0

../4

The alignement can be made in the same way as described for the output stage.

Dummy load: 240 ohms in parallel with 3.3 pF.

Temporary class-A adjustment at I_C = 44mA.

5. OVERALL PERFORMANCE

The minimum performance to be expected is:

Drive power: 55 mW; Power gain: 21.6dB

Current consumption: 1.63A; Efficiency: 39.3%

Typically the results can be:

Drive power: 25mW; Power gain: 25 dB

Current consumption : 1.33A; Efficiency: 48%.

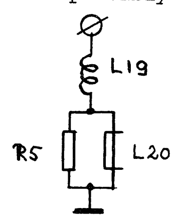
6. OUTPUT POWER CONTROL

If only a few dB of power control is needed it can be achieved by varying V_{S1} only.

For control over a larger range e.g. tens of dB's it is recommended to vary both V_{S1} and V_{S2}.

7. STABILITY

If parasitic oscillations occur during load mismatch it can be useful to replace L14 and possibly L8 by the following network:



R5 = 10Ω, carbon

L19 = 100 nH; 7 turns closely wound enamelled Cu wire (0,5 mm); int. diam.3 mm; leads 2x5 mm.

L20 = L3 (See main parts list).

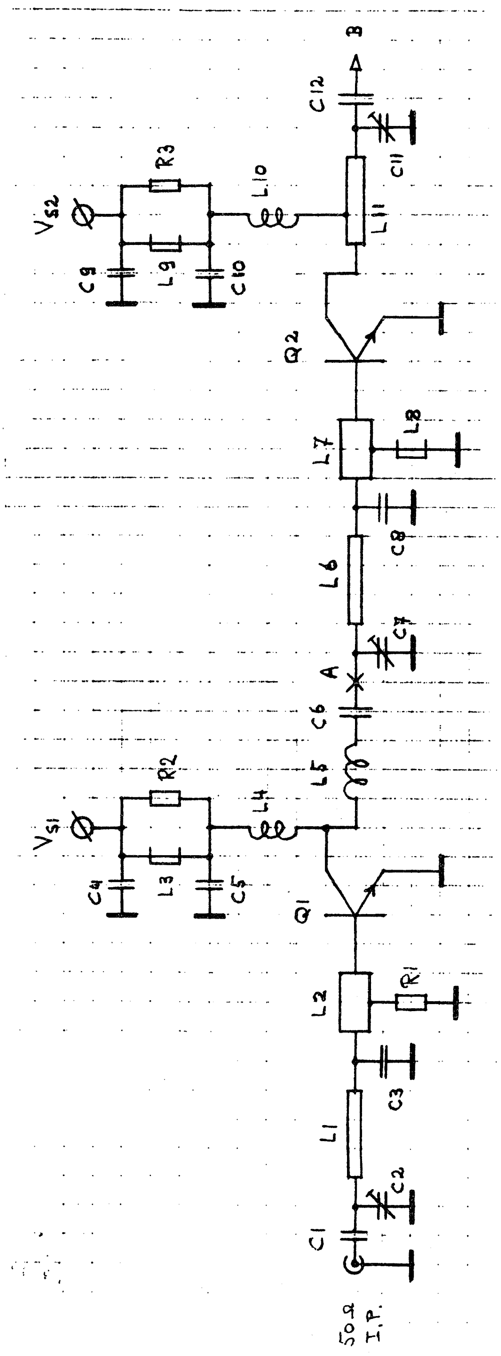
8. P.C. BOARD MATERIAL

The stripline calculations in this report are based on PTFE fibre-glass material with E_r = 2.74, dielectric thickness of 718μm and copper thickness of 38.1μm.

If another material is used it can be useful to know the "free-air" length of the striplines in this design. For this purpose the lengths given in the parts list must be multiplied by the so-called shortening factor which is given in the table below.

$R_c (\Omega)$	W(mm)	Shortening factor
71.7	1.0	1.455
57.1	1.5	1.476
36.0	3.0	1.516

A.H.Hilbers



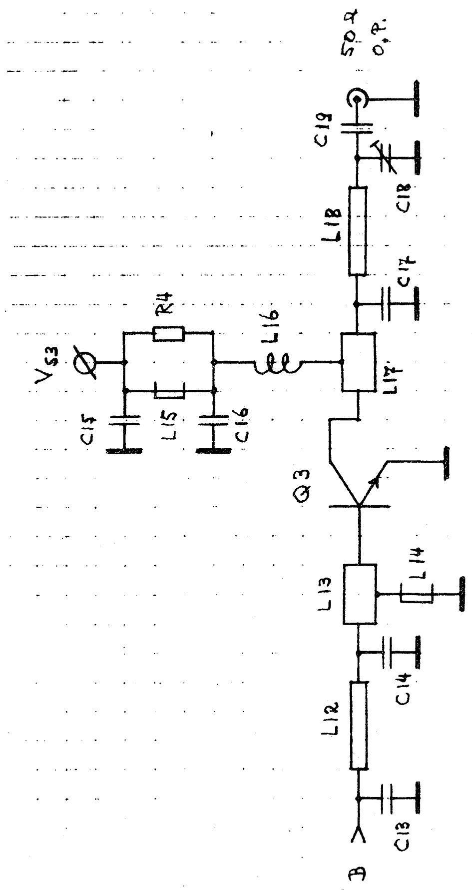


Fig.1

820 - 870 MHz power amplifier

PARTS LIST

Q1 = BLU 98
Q2 = BLV 91
Q3 = BLV 93

R1 = 270 Ω, carbon
R2 = 10 Ω, carbon
R3 = 10 Ω, carbon
R4 = 10 Ω, carbon

C1 = 27pF, chip
C2 = parallel connection of 3,3pF, chip and 1.2-3.5pF PTFE trimmer [1)]
C3 = 2x 10pF, chips in parallel
C4 = 100 nF, polyester
C5 = 27pF, chip
C6 = 27pF, chip
C7 = parallel connection of 2.7pF, chip and 1.2-3.5pF PTFE trimmer.
C8 = 2x 8.2pF, chips in parallel
C9 = 100 nF, polyester
C10 = 27pF, chip
C11 = 1.2-3.5pF PTFE trimmer
C12 = 27pF, chip
C13 = 5.6pF, chip
C14 = 2x10pF, chips in parallel
C15 = 100 nF, polyester
C16 = 27pF, chip
C17 = 2x6.8pF, chips in parallel
C18 = parallel connection of 1.8pF, chip and 1.2-3.5pF PTFE trimmer
C19 = 27pF, chip

L1 = 71.7Ω stripline; W= 1.0 mm, l= 14.0 mm
L2 = 36Ω stripline; W= 3.0 mm, l= 4.0 mm
L3 = FXC-bead, grade 3S1, cat.nr. 4322 030 32160,
wound with 4 turns of 0,5 mm enamelled Cu wire.
L4 = 15.1nH; 2 turns of 0,5 mm Cu wire on d=2,5 mm, l = 2,3 mm
leads: 2x5mm
L5 = 16.3 nH; 2 turns of 0.5 mm Cu wire on d= 2.5 mm, l= 1.9 mm,
leads: 2x5 mm
L6 = 71.7Ω stripline; W= 1.0 mm, l= 15.2 mm
L7 = 36Ω stripline; W= 3.0 mm, l= 4.2 mm
L8 = L3

L10 = 52nH; 6 turns of 0,5 mm enamelled Cu wire on d= 2.0 mm, l= 3.6 mm, leads: 2x5 mm, connected to L11 at 2.0 mm from the edge of the transistor

L11 = 71.7Ω stripline; W= 1.0mm, l= 22.7 mm

L12 = 71.7Ω stripline; W= 1.0mm, l= 13.9mm

L13 = 36Ω stripline; W= 3.0mm, l= 3,3mm

L14 = L3

L15 = L3

L16 = 15nH; 2 turns of 0.5mm Cu wire on d= 2,5mm, l= 2,3mm leads: 2x5mm, connected to L17 at 4.0mm from the edge of the transistor

L17 = 36Ω stripline; W= 3.0mm, l= 10.1mm

L18 = 57.1Ω stripline; W=1.5 mm, l= 21.6mm

P.c.board material: 1/32" PTFE fibre-glass, E_r=2.74
as delivered by PERMALI Ltd, Gloucester, U.K.

1) cat.nr. 2222 809 05001.

N.V. PHILIPS SEMICONDUCTORS NIJMEGEN – THE NETHERLANDS

PRODUCT GROUP SPECIALTIES AND DIODES

Report no.: NCO 8404 RNR-1-367-1984-AS
Author : A.H.Hilbers
Date : 1984-10-01

APPLICATION

A 3-STAGE LINE-UP FOR THE 900MHz BAND WITH 8W OUTPUT POWER AT 12.5V

ABSTRACT

A 3-stage power amplifier for mobile radio transmitters in the frequency band from 880 to 925MHz has been described.
The output power is 8W at a supply voltage of 12.5V.
The amplifier uses the following transistor types:
BLU 98, BLV 91 and BLV 93.
The required drive power is 65mW max. and the overall efficiency is better than 39%.

1. INTRODUCTION

In our Semiconductor Handbook, part S6, 1984 we give a number of line-ups for mobile radio transmitters in the 800-960MHz band.
They are intended for a supply voltage of 12.5V.
The first one of these line-ups with an output power of 8W is theoretically investigated for the frequencyband of 880-925MHz and is described in this report.
The individual stages have been designed on basis of 50 ohms input and load impedances. In the schematic diagram of Fig.1 the intermediate 50 ohm points have been indicated by the characters A and B.
In this way the stages can be tested and aligned individually.

2. THE OUTPUT STAGE

This stage has been built around the BLV 93 which is a JO common-emitter transistor in a 6-leads flange package with rectangular ceramic cap (SOT 171).
The transistor is specified at 900MHz in class-B for 8W output power at 12.5V:
Power gain: 6.5dB min., 7.3dB typ.
Drive power: 1.8W max., 1.5W typ.
Efficiency : 50% min., 58% typ.
Collector current: 1.28A max., 1.1A typ.

../2

The input and optimum load impedance in this frequency band are given in the table below:

f (MHz)	Z_i (Ohms)	Z_L (Ohms)
880	3.36 + j 2.30	6.25 + j 0.36
900	3.50 + j 2.22	6.08 + j 0.29
925	3.68 + j 2.05	5.84 + j 0.17

The circuit has been designed for 1/32" PTFE fibre-glass with a relative dielectric constant of 2.74 (See parts list).

For the alignment of the output network it can be useful to replace the transistor by a dummy load which is connected between the collector and emitter terminals.
This load must consist of the parallel connection of an 8.2 Ohm chip resistor and a 12pF chip capacitor.
The output circuit must be aligned such that the VSWR measured at the 50 Ohms output terminal is minimum.
To align the input circuit the BLV 93 can be temporarily adjusted in class-A at I_c=1.1A. Because of the built-in emitter resistance the transistor can withstand this for a longer time provided it is mounted on an adequate heatsink. The class-A biasing can be achieved by disconnecting L14 at the ground side and by applying a small positive voltage at this point, which must of course be well decoupled.
The input VSWR can now be measured at point B on a small signal basis and the circuit must be aligned to minimize this VSWR.
This method is intended as a coarse alignment. Final alignment must be done in class-B at the actual power level.
In general it can be said that both value and position of C14 and C17 are rather critical.
C17 controls the efficiency over the band and C14 the input VSWR.

3. THE DRIVER STAGE

This stage uses the BLV 91, a common emitter transistor in a 4-leads stud package with ceramic cap (SOT 172).
The transistor is used in class-B with a maximum output power of 1.8W at 12.5V.
Under these conditions and at 925MHz we can expect:
Power gain : 6.5dB min., 7.8dB typ.
Drive power : 0.4W max., 0.3W typ.

../3

Efficiency: 50% min., 60% typ.

Collector current: 0.29A max., 0.24A typ.

The input and optimum load impedance are given in the table below:

f(MHz)	Z_i (Ohms)	Z_L (Ohms)
880	4.04 + j 3.25	18.6 + j 15.9
900	4.04 + j 3.52	18.7 + j 16.0
925	4.03 + j 3.84	18.8 + j 16.1

The alignment of this stage can be done in the same way as that of the output stage. For the dummy load a parallel connection of 39 Ohms and 4.7pF is recommended and for temporary class-A adjustment a collector current of 0.24A can be chosen. A 50 Ohm source must be connected to point A via a blocking capacitor (24pF, chip).

4. THE PRE-DRIVER STAGE

In this stage a BLU 98 is used. This is a common-emitter transistor in a 4-leads plastic pill package (SOT 103).
It is used in class-B with a maximum output power of 400mW at 12.5V.
Under these conditions and at 925MHz we can expect:

Power gain : 8.0dB min., 9.5dB typ.
Drive power : 63mW max., 45mW typ.
Efficiency : 50% min., 65% typ.
Collector current: 64mA max., 49mA typ.

The input and optimum load impedance are given in the table below:

f(MHz)	Z_i (Ohms)	Z_L (Ohms)
880	3.28 + j 2.40	12.0 + j 44.2
900	3.39 + j 2.83	12.1 + j 44.4
925	3.54 + j 3.37	12.3 + j 44.9

The alignment can be made in the same way as described for the output stage.
Dummy load: 220 Ohms in parallel with 3.3pF.
Temporary class-A adjustment at I_C=49mA.

../4

5. <u>OVERALL PERFORMANCE</u>

The minimum performance to be expected is:

Drive power : 65mW; Power gain : 21.0dB

Current consumption: 1.63A; Efficiency: 39.3%

Typically the results can be:

Drive power: 28mW; Power gain: 24.5dB

Current consumption: 1.33A; Efficiency: 48%.

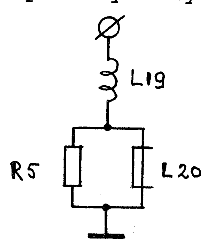
6. <u>OUTPUT POWER CONTROL</u>

If only a few dB of power control is needed it can be achieved by varying V_{S1} only.

For control over a larger range e.g. tens of dB's it is recommended to vary both V_{S1} and V_{S2}.

7. <u>STABILITY</u>

If parasitic oscillations occur during load mismatch it can be useful to replace L14 and possibly L8 by the following network:



R5 = 10Ω , carbon

L19= 100nH; 7 turns closely wound enamelled Cu-wire (0.5mm);
int.diam. 3mm; leads 2x5mm.

L20= L3 (See main parts list)

8. <u>P.C.BOARD MATERIAL</u>

The stripline calculations in this report are based on PTFE fibre-glass material with E_r=2.74, dielectric thickness of 718μm and copper thickness of 38.1μm.

If another material is used it can be useful to know the "free-air" length of the striplines in this design.

For this purpose the lengths given in the parts list must be multiplied by the so-called shortening factor which is given in the table on next page.

$R_c (\Omega)$	W(mm)	Shortening factor
71.7	1.0	1.455
57.1	1.5	1.476
36.0	3.0	1.516

A.H.Hilbers

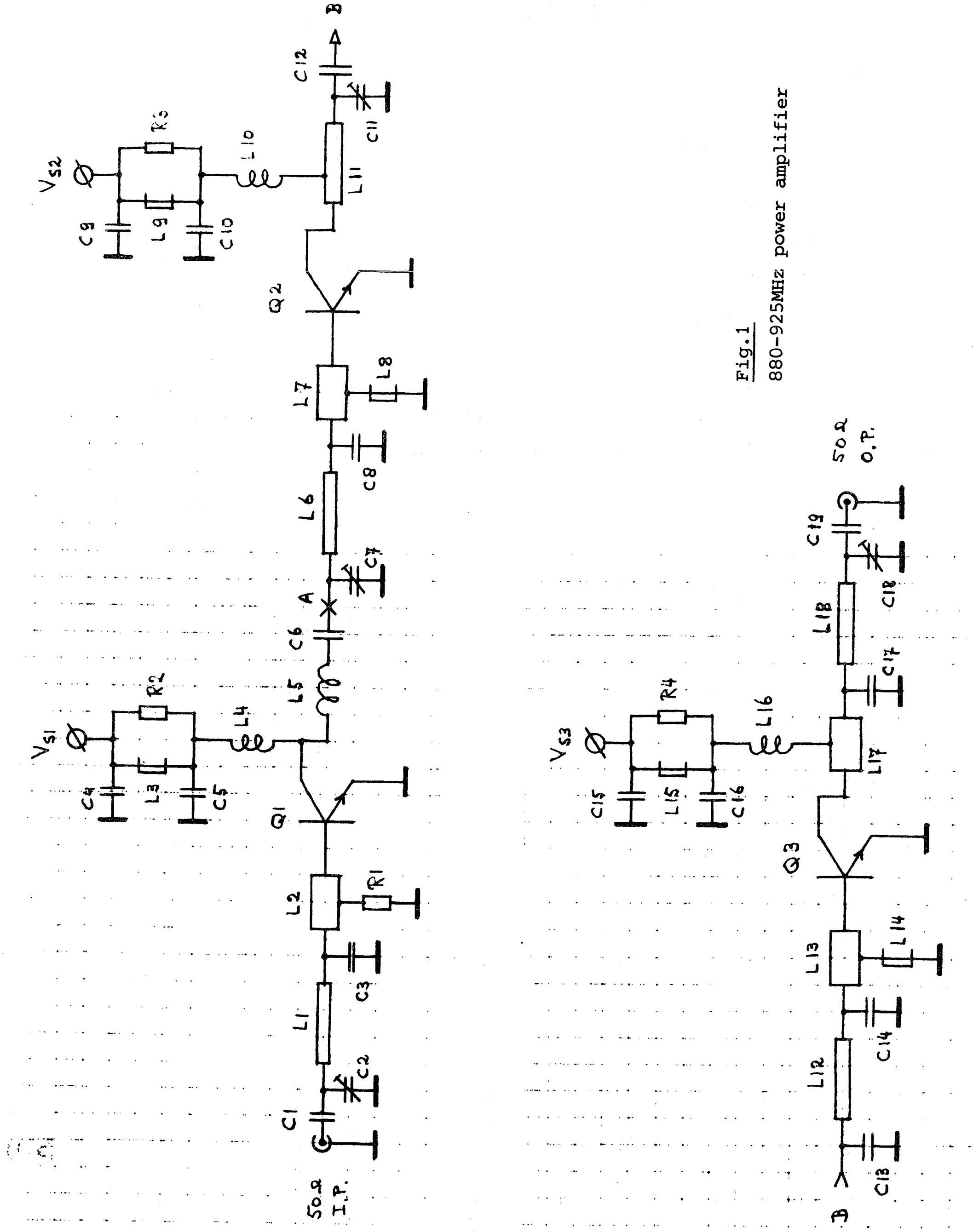


Fig.1
880-925MHz power amplifier

PARTS LIST

Q1 = BLU 98
Q2 = BLV 91
Q3 = BLV 93

R1 = 270Ω , carbon
R2 = 10Ω , carbon
R3 = 10Ω , carbon
R4 = 10Ω , carbon

C1 = 24pF, chip
C2 = parallel connection of 2.4pF, chip and 1.2-3.5pF PTFE trimmer[1)]
C3 = parallel connection of 9.1pF and 8.2pF, chips.
C4 = 100nF, polyester
C5 = 24pF, chip
C6 = 24pF, chip
C7 = parallel connection of 2.2pF, chip and 1.2-3.5pF PTFE trimmer.
C8 = parallel connection of 8.2pF and 7.5pF, chips.
C9 = 100nF, polyester
C10= 24pF, chip
C11= 1.2-3.5pF PTFE trimmer
C12= 24pF, chip
C13= 4.7pF, chip
C14= 2x8.2pF, chips in parallel
C15= 100nF, polyester
C16= 24pF, chip
C17= 2x6.2pF, chips in parallel
C18= parallel connection of 1.8pF, chip and 1.2-3.5pF PTFE trimmer
C19= 24pF, chip

L1 = 71.7Ω stripline; W=1.0mm, l=13.6mm
L2 = 36Ω stripline; W=3.0mm, l=3.1mm
L3 = FXC-bead, grade 3S1, cat.nr. 4322 030 32160, wound with 4 turns
of 0.5mm enamelled Cu-wire.
L4 = 14.9nH; 2 turns of 0.5mm Cu-wire on d=2.5mm, l=2.4mm, leads: 2x5mm
L5 = 13.9nH; 2 turns of 0.5mm Cu-wire on d= 2.5mm, l=2.8mm, leads:2x5mm
L6 = 71.7Ω stripline; W=1.0mm, l=14.2mm
L7 = 36Ω stripline; W=3.0mm, l=3.1mm
L8 = L3

L9 = L3

L10= 52nH; 6 turns of 0.5mm enamelled Cu-wire on d=2.0mm,l=3.6mm,leads:2x5mm, connected to L11 at 2.0mm from the edge of the transistor.

L11= 71.7Ω stripline; W=1.0mm, l=21.5mm

L12= 71.7Ω stripline; W=1.0mm, l=13.4mm

L13= 36Ω stripline; W=3.0mm, l=3.8mm

L14= L3

L15= L3

L16= 15nH; 2 turns of 0.5mm Cu-wire on d=2.5mm, l=2.3mm,leads: 2x5mm, connected to L17 at 4.0mm from the edge of the transistor.

L17= 36Ω stripline; W=3.0mm, l=8.3mm

L18= 57.1Ω stripline; W=1.5mm, l=19.6mm

P.C.board material: 1/32" PTFE fibre-glass, E_r=2.74 as delivered by:
PERMALI Ltd.
Gloucester, U.K.

1) Cat.nr. 2222 809 05001

PHILIPS | PRODUCT GROUP SPECIALTIES AND DIODES NIJMEGEN | APPLICATION

Report no : RNR-1-387-1987-AS / NCO 8706
Author : H.van Hees
Date : 1987-06-30

A WIDEBAND AMPLIFIER (890-915MHz) WITH THE TRANSISTORS BLV92 AND BLV94

SUMMARY

For mobile radio transmitters in the frequency band 890-915MHz a class B power amplifier has been designed around the transistor BLV92 and BLV94. The BLV92 is used in the common emitter and the BLV94 in common base configuration. The applied matching networks are of the low-pass Chebychev type.

The main amplifier properties are:

Frequency range		: 890-915MHz
Supply voltage	V_S	: 12.5V
Drive power	P_{dr}	: 400mW
Load power	P_L	: >14W
Input return losses		: >15dB
Efficiency		: >45%
Load and source impedance		: 50Ω

Applied PC board material is double copper clad teflon fibre glass (ε_r=2.74), thickness 1/32".

INTRODUCTION

The BLV92 and BLV94 are transistors in a SOT171 envelope, for application in mobile radio transmitters in the 900MHz band. A wideband amplifier has been designed around these transistors, for the frequency range 890-915MHz. The individual stages have been designed on basis of 50Ω input and load impedances. The interstage impedance is 50Ω.

OUTPUT STAGE

This stage is a design around the BLV94, which is a common-base transistor with internal input matching in a 6 leads flange envelope with a ceramic cap (SOT171). The transistor is specified at 900MHz in class B, for 15W output power at 12.5V supply voltage:

power gain	:	min. 6dB ,	typical 7 dB
drive power	:	max. 3.75W,	typical 3 Watt
efficiency	:	>50% ,	typical 60%
current	:	max. 2.4A ,	typical 2A

The input- and optimum load impedance are given below.

Freq.(MHz)	Zi (Ω)	Zload (Ω)
880	4.0 + j 5.50	1.75 - j 1.5
890	4.0 + j 5.25	1.80 - j 1.5
900	4.0 + j 5.00	1.8 - j 1.6
915	3.8 + j 4.75	2.1 - j 1.8
920	3.75+ j 4.70	2.2 - j 2.0

INPUT STAGE

This stage is a design around the BLV92, which is a common-emitter transistor with internal input matching, in a SOT171 header.
The transistor is specified at 900MHz in class-B for 4W output power at 12.5V supply voltage:

power gain	: min. 7.5dB	,	typical 8.5dB
drive power	: max. 710mW	,	typical 565mW
efficiency	: >50%	,	typical 57%
current	: max.640mA	,	typical 560mA

The input- and optimum load impedances are given below.

Freq.(MHz)	Zi (Ω)	Zload (Ω)
880	5.5 + j 3.2	9 + j 4.0
890	6.0 + j 3.0	8.8 + j 3.9
900	6.2 + j 3.0	8.6 + j 3.8
915	6.4 + j 2.5	8.5 + j 3.6
920	6.5 + j 2.5	8.4 + j 3.6

CIRCUIT DESIGN

With the impedance values and CAD programmes we have designed a stripline circuit on teflon PC board (ε_r=2.7), thickness 1/32". Page 4 shows the electrical circuit. The impedance transformations have been established by means of low-pass Chebychev filter designs. The intermediate impedance between the two stages is 50Ω.
The 3B-beads, the R2/L13 resp. R1/L14 combinations and R3 are necessary components to establish stable behaviour during mismatch conditions.

Page 5 shows the PC board. Actual dimensions 60' x45mm.
Fig. 3 on page 5 shows the amplifier lay-out. At the edges and at the emitter earthing points the upper and lower side of the PC board have been connected by means of copper straps to improve earthing. For that purpose also rivets have been applied at several places. The PC board and transistors are screwed onto a copper intermediate heatsink, thickness 10mm. When operating the amplifier this heatsink has to be attached to a larger heatsink in order to keep transistor temperature at an acceptable level.

AMPLIFIER PERFORMANCE

The table below summarizes the amplifier performance.
Conditions: V_S=12.5V, P_{drive}=400mW, T_H=25^{o}C.

Freq.range	:	890-915MHz
Outputpower P_L	:	>14.4 Watt
Efficiency	:	>46%
Amplifier current	:	<2.5A
Input VSWR	:	<1.4:1

Fig.4, page 7, shows the ouput power, efficiency and input return losses versus frequency at V_S=12.5 Volt and P_{drive}=400mW.

Fig.5 page 8 gives P_L versus P_{drive} and gain, efficiency and input return losses versus P_L at 890MHz and 915MHz.

Harmonic content
P_{drive}=400mW, V_S=12.5 Volt.

fo=0dB	2fo	3fo
890MHz	-44dB	-50dB
915MHz	-49dB	-55dB

Stability
All non harmonically related output signals are below -50dB under the following conditions:
- freq. 890-915MHz
- V_S= 5-15.5 Volt
- P_{drive}= 100-600mW
- VSWR= 3:1 trough all phases.

Ruggedness
The amplifier is able to withstand an output mismatch of 50:1 through all phases at any frequency in the band at a supply voltage of V_S=15.5 Volt and a drive power of P_{dr}=400mW.

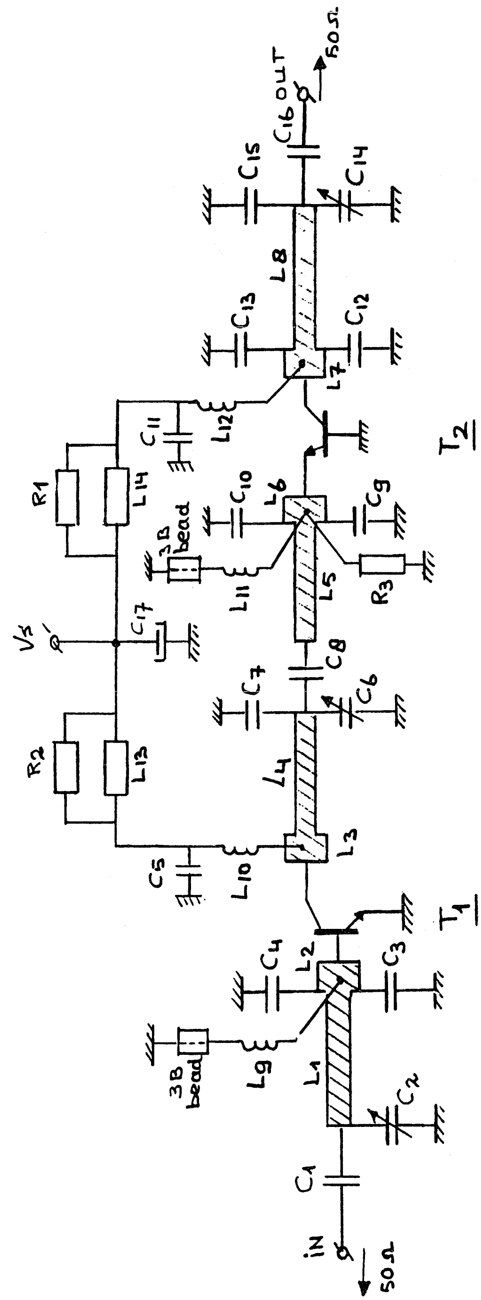


Fig. 1. Electrical circuit.

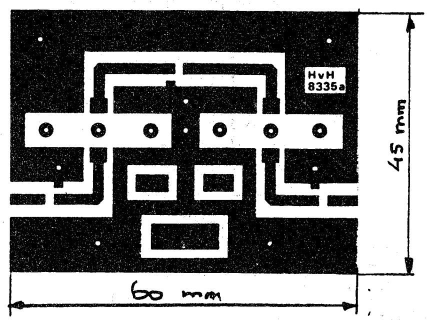


Fig. 2. p.c. board.

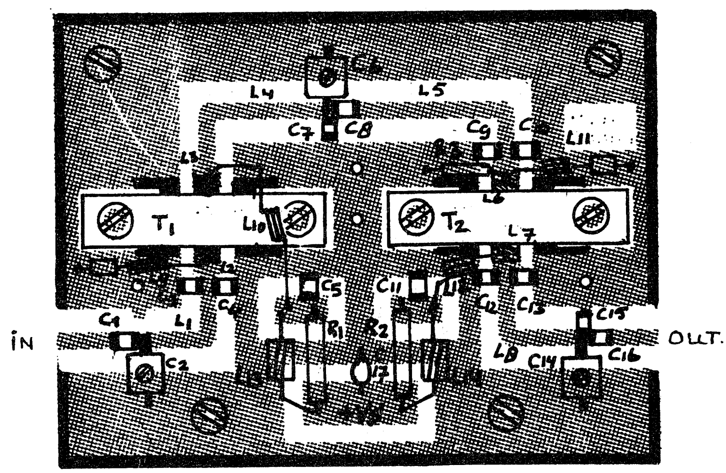
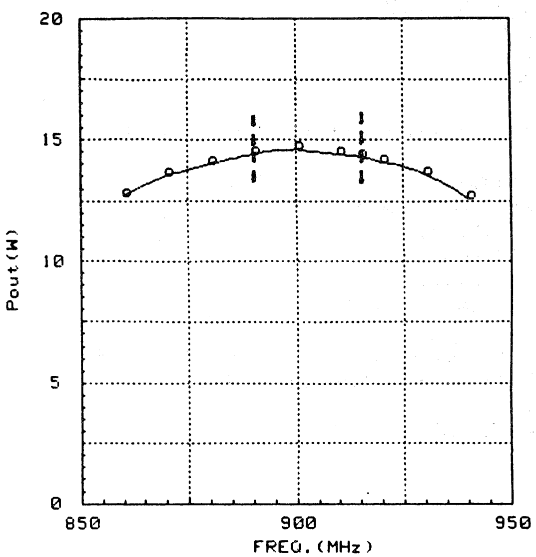


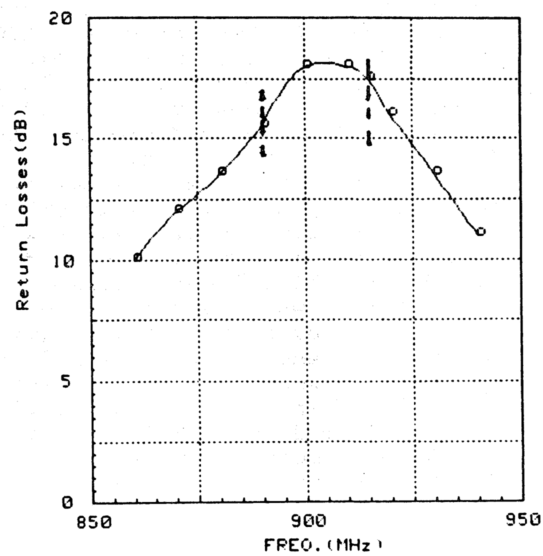
Fig. 3. Amplifier lay-out.

Parts List

C1=C5=C8=C11=C16=33pF chip capacitor. ATC 100A
C3=C4=C7=6.2pF chip capacitor. ATC 100A
C9=C10=7.5pF chip capacitor. ATC 100A
C12=C13=10pF chip capacitor. ATC 100A
C15=3.9pF chip capacitor. ATC 100A
C2=C6=C14=1-3.5 film dielectric trimmer,Pilips cat nr.2222.809.05001
C17=2.2uF 35V tantalium.
L1=50Ω stripline,length 12mm,width 1.86mm.
L2=L3=L6=L7=36Ω stripline,length 2.5mm,width 3mm.
L4=50Ω stripline,length 17.7mm,width 1.86mm.
L5=50Ω stripline,length 20mm,width 1.86mm.
L8=50Ω stripline,length 13.2mm,width 1.86mm.
L9=L10=L11=4 wnd.enamelled Cu-wire,ø 0.4mm,closely
wound,int.diam.D=3.5mm.
L12=5 wnd.enamelled Cu-wire,ø 0.4mm,closely wound,int.diam.D=4.5mm.
L13=L14=2.5 wnd.enamelled Cu-wire,ø 0.4mm,on a ferrite 3B-bead,
Philips cat.nr.4330.030.32261.
Bead=ferrite 3B-bead.Philips cat nr.4330.030.32221.
R1=R2=10ohm metal film resistor.
R3=75 ohm metal film resistor.
T1=BLV92.
T2=BLV94.

pc-board material is double copper clad PTFE fibre glass,
thickness 1/32 inch,Er=2.74.





Pdr.=400mW/Vs=12.5V.

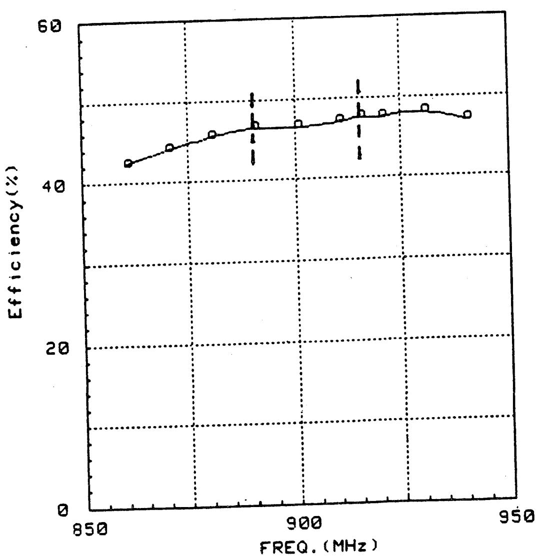
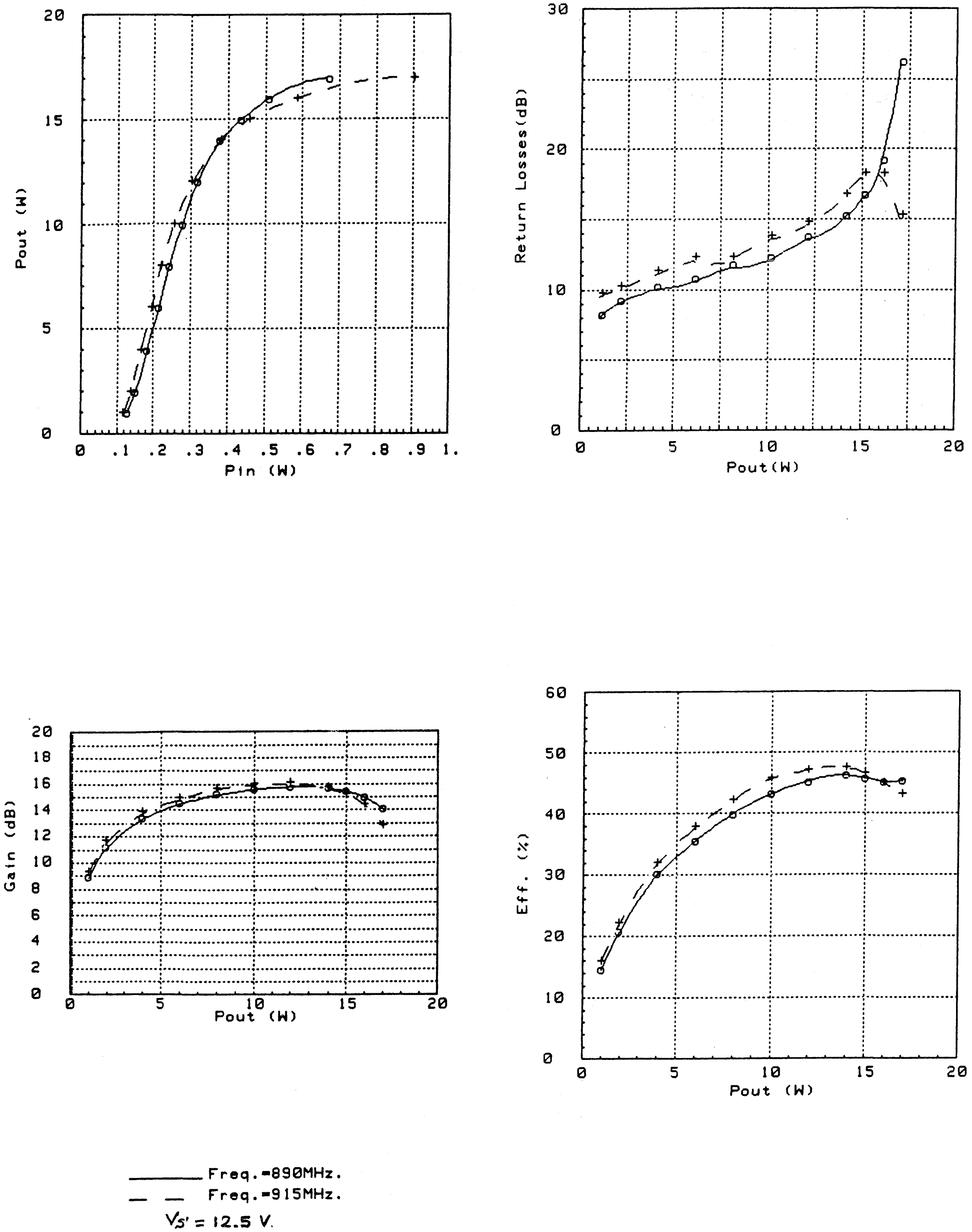


Fig. 4.



Pout (W)
Pin (W)
Return Losses(dB)
Pout(W)
Gain (dB)
Pout (W)
Eff. (%)
Pout (W)
Freq.=890MHz.
Freq.=915MHz.
$V_{S'}$ = 12.5 V.

Fig. 5.

Report no : RNR-1-428-1987-AS / NCO 8707
Author : M.J.Köppen
Date : 1987-08-17

TITLE

A WIDEBAND 4 STAGE BASE-STATION POWER AMPLIFIER FOR 900MHz CELLULAR RADIO SYSTEMS.

SUMMARY

In this report a brief description is given of a wideband power amplifier for the 900MHz band.
It utilizes two common-base circuited BLV97 transistors, coupled by means of 3dB-90^{o} hybrids, in the final stage.
The nominal output power amounts to 50W.

The complete amplifier contains 4 stages in cascade, viz. single-ended drivers BLV98-BLV99 operating in class-B and BLU98 in class-AB.

Being tuned for the (original) PRCS-band of 937-941MHz the typical overall gain is 36.4dB, whilst the typical efficiency amounts to 49%. These values are valid for P_L=50W; V_S=24V; input and output impedances of 50Ω.

The amplifier has been designed for a tuning range of at least 870-950MHz.

Applied pc board material is double Cu clad with PTFE fibre-glass dielectric (ε_r=2.2); thickness 1/32 inch.

INTRODUCTION

The BLV97 and BLV98 are transistors in SOT171 envelopes, whilst the BLV99 is housing in an SOT172. The predriver BLU98 is encapsulated in a subminiature plastic transfer-moulded cross-package SOT103 and operates in class-AB from 12V supply voltage via an μA7812 voltage stabilizer. Supply voltage for the complete amplifier module amounts to 24V.

Above mentioned devices are specially designed for use in the 900MHz communications band.

To obtain at least 50W output power a pair of BLV97's is required. Under worst-case conditions they ask for a drive power of approx. 14W, what can be done with a single BLV98. Preceding stages are BLV99 for at least 2W output and the smaller BLU98 being published for 0.5W.

OUTPUT STAGE

This stage is a design around a pair of BLV97's, which are common-base transistors with internal pré-matching in an 6-leads flange envelop with ceramic cap.

Fig.1 shows the principle of the hybrid coupled output stage.

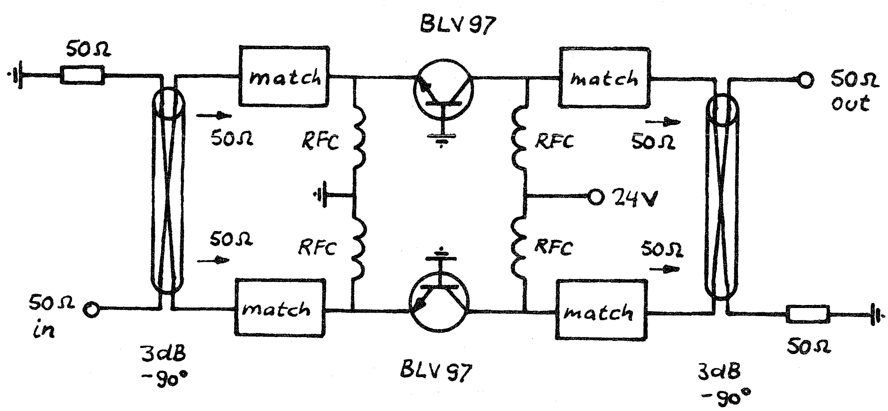


Fig.1

The hybrids are constructed with suitable lengths of SAGE WIRELINE. It has been chosen for the semi rigid (seamless copper tube) version BHC (electrical length ¼ λ ; physical length 52mm for the 900MHz band).

Hybrid resistors R9-14 (see circuit diagram Fig.2A/B) are power types (20W) on BeO.
They are screwed to the intermediate heatsink base plate.

For wideband input and output matching in principle the Chebishev low-pass response has been chosen. Afterwards some optimization has been applied with the aid of a CAD programme. Because of this modification and introducing some tuning provisions (trimmers), it may be expected that the originally chosen Chebishev response is somewhat deformed, but still acceptable.

For transforming sections mainly strip transmission line techniques have been applied.

The characteristic impedances of the striplines around the transistors are mainly appointed by the lead-width of the devices (Z_C=39Ω).

Ferroxcube leads L36-39-46-49, resistors R10-11-12-13, inductors L37-38-47-48 and decoupling capacitors C52-53-54-55-64-65 are necessary components to establish stable behaviour during mismatch conditions for devices operating on a 6dB/octave slope. (See complete circuit diagram Fig. 2A/B).

ELECTRO-MECHANICAL DETAILS

The final stage takes about half the total surface of the pc board.Actual dimension of the complete pc board, including driver stages, amounts to 252mm x 80mm. Fig.3 shows the copy of the positive, whilst in Fig.4 the complete amplifier lay-out has been sketched.

To keep dielectric losses as low as possible and for universality it has been chosen for RT-DUROID type 5880 (ε_r=2.20) double clad board with PTFE fibre-glass dielectric (thickness 1/32 inch; Cu sheet 2x17μm).

Because plating-through could not been realized in the prototypes (4 units), it has been imitated by means of soldering groups of gold-plated subassemblies of semiconductor diodes. These components came from our diode fabrication department.

At the edges, near both coaxial plugs and the emitter earthing points of the pre-driver stages some copper straps are connected between upper and lower side of the board to improve earthing.
For that purpose rivets have been applied (and soldered) at some places too.

The pc board, transistors and hybrid resistors are screwed onto an aluminia intermediate heatsink, thickness 10mm. When operating the amplifier module this base-plate has to be attached to a larger heatsink in order to keep transistor temperatures at an acceptable level.

Fig.5 shows the drawing of the base-plate. To prevent feedback all wiring (PTFE isolated) has been led through the fraised channels in the base-plate. For that purpose the lower side of the pc board needs some holes or isolated contact areas.

DRIVER STAGES

Following the foregoing it will be clear in what way the three preceding stages were created. Here the method was the same as described for the final stage.

For space-saving several striplines have been bent (in meanders) however preventing parasitic coupling as much as possible.

The 12V stabilizer and the multiturn potentiometer for the I_{CZS} setting (approx.3mA) of the BLU98 are attached to the inner side of the raised border, being screwed to the far end of the base-plate.

Both raised borders (input and output) contain the coaxial N-chassis plugs (male and female).
Sizes of the borders are 82mm x 40mm x 2.5mm. The unit further contains an U-shaped metal cover, being screwed to the base-plate, to close it.

It should be noted that some detuning is possible when the module is closed. That problem can be solved by applying a dummy cover or an exchangable cover with tuning holes.

Diode D1, being applied for temperature stabilization of the BLU98 bias point, has to make thermal contact with the plastic body of T1. (Advise: apply some heatsink compound).

AMPLIFIER PERFORMANCE

Fig.6 gives a survey of the practical results.
In this case the amplifier was tuned for the upper-band segment of 937-941MHz (2 test units have been checked).

Because the design was set-up for the complete range of 870-950MHz it is possible to tune the amplifier for other bandsegments (up to 20MHz bandwidth).

Practical experiments have been made with an 870-890MHz segment (2 units).

Typical results in figures were:
870MHz: G_T = 40.3dB : eff_T = 50.3%
880MHz: G_T = 41.0dB : eff_T = 52.6%
890MHz: G_T = 41.0dB : eff_T = 53.1%

No output mismatch tests have been done; however no problems were indicated during all tuning procedures.

All non-harmonically related output signals (spurious) were at least -60dB vs. carrier.

PRCS = Personal Radio Communication System.

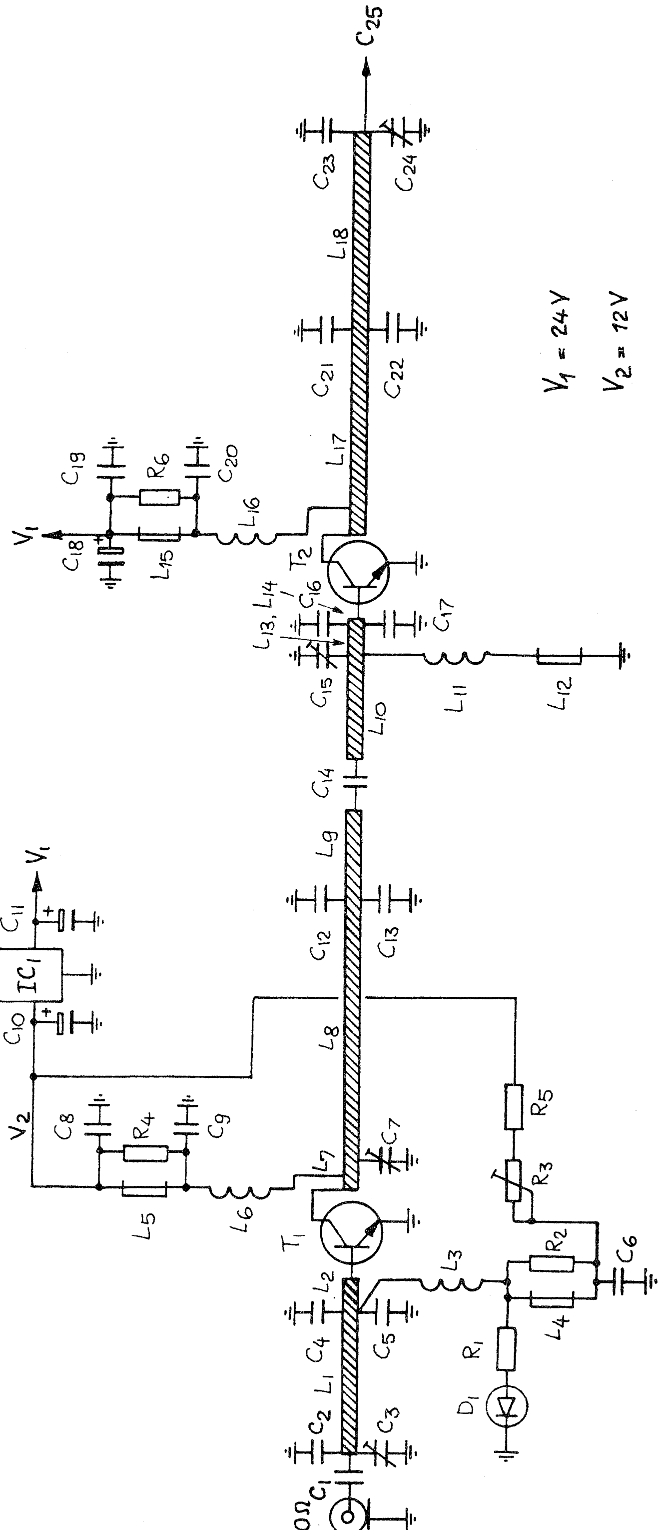


FIG.2A

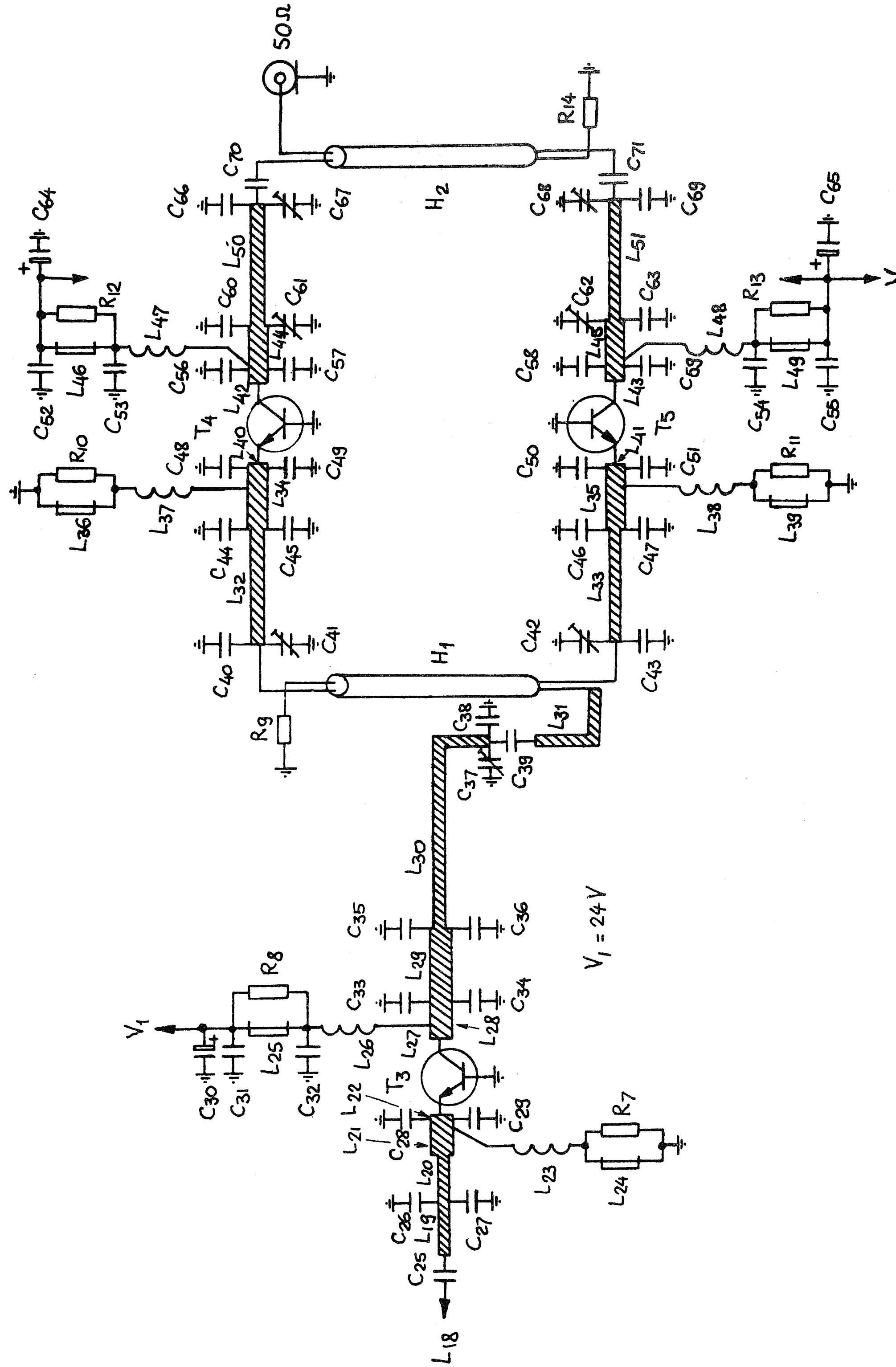


FIG.2B

LIST OF COMPONENTS

C1=C14=C25=C39=47pF ± 5% (50V)	(2222 852 12479)*
C2=1pF ± 5% (50V)	(2222 851 12108)*
C3=C7=C15=C24=C37=C41=C42=C61=C62=C67=C68= 1.2 to 3.5pF film dielectric trimmer	(2222 809 05001)
C4=C5=C44=C45=C46=C47=8.2pF ± 5% (50V)	(2222 851 12828)*
C6=C52=C55=10nF ± 10% (50V)	(2222 852 47103)*
C8=100nF ± 20% (50V)	(2222 852 48104)*
C9=C19=C20=C31=C32=C53=C54=100pF ± 10% (50V)	(2222 852 13101)*
C10=C11=1µF (35V) tantalum electrolytic capacitor	
C12=C13=2.2pF ± 5% (50V)	(2222 851 12129)*
C16=C17=6.8pF* **	
C18=C30=C64=C65=10µF (63V) electrolytic capacitor	
C21=C27=4.7pF ±5% (50V)	(2222 851 12478)*
C22=C35=C36=5.6pF ± 5% (50V)	(2222 851 12568)*
C23=C38=1.5pF ± 5% (50V)	(2222 851 12158)*
C26=C40=C43=3.9pF ± 5% (50V)	(2222 851 12398)*
C28=C29=C33=C34=12pF ± 5% (50V)	(2222 851 12129)*
C48=C49=C50=C51=10pF ± 5% (50V)	(2222 851 12109)*
C56=C57=C58=C59=18pF* **	
C60=C63=5.6pF* ***	
C66=C69=0.5pF* ***	
C70=C71=100pF* ***	

*   Multilayer ceramic chip capacitor
** American Technical Ceramics capacitor type 100A or capacitor of same quality
*** Idem, type 100B

LIST OF COMPONENTS CONTINUED

R1=12Ω ± 2%; 0.25W metal film resistor
R2=R4=R6=R8=10Ω ± 10%; 0.25W metal film resistor
R3=1kΩ multiturn potentio-meter
R5=1.2KΩ ± 10%; 0.25W metal film resistor
R7=R10=R11=1Ω ± 10%; 0.25W metal film resistor
R9=R14=50Ω; power resistor (20W) on BeO
(Pyrofilm Corp. type PPR 515-20-3)
R12=R13=10Ω ± 10%; 1W metal film resistor

L1=50Ω stripline (24.0mm x 2.39mm)
L2=50Ω stripline (5.5mm x 2.39mm)
L3=L6=L11=L16=83nH; 9 turns closely wound enamelled
Cu wire (0.4mm); int.dia.2mm; leads approx.2x4mm
L4=L5=L12=L15=L24=L36=L39=Ferroxcube wideband hf choke,
grade 3B (4312 030 32261) with 3 turns enamelled
Cu wire (0.4mm)
L7=50Ω stripline (7mm x 2.39mm)
L8=50Ω stripline (42.6mm x 2.39mm)
L9=50Ω stripline (15.3mm x 2.39mm)
L10=50Ω stripline (18.1mm x 2.39mm)
L13=50Ω stripline (5.0mm x 2.39mm)
L14=50Ω stripline (1.0mm x 2.39mm)
L17=50Ω stripline (34.7mm x 2.39mm)
L18=50Ω stripline (32.9mm x 2.39mm)
L19=50Ω stripline (9.6mm x 2.39mm)
L20=50Ω stripline (9.0mm x 2.39mm)
L21=39Ω stripline (6.1mm x 3.42mm)
L22=39Ω stripline (1.0mm x 3.42mm)
L23=L37=L38=250nH; 13 turns closely wound enamelled
Cu wire (0.4mm); int.dia. 3mm; leads approx.2x3mm

LIST OF COMPONENTS CONTINUED

L26=L47=L48=6 turns closely wound enamelled
Cu wire (0.7mm); int.dia. 3mm; leads approx.2x5mm
L27=39Ω stripline (4.5mm x 3.42mm(
L28=39Ω stripline (2.8mm x 3.42mm)
L29=39Ω stripline (14.2mm x 3.42mm)
L30=50Ω stripline (31mm x 2.39mm)
L31=50Ω stripline (28mm x 2.39mm)
L32=L33=50Ω stripline (21.5mm x 2.39mm)
L34=L35=39Ω stripline (11.0mm x 3.42mm)
L40=L41=39Ω stripline (1.2mm x 3.42mm)
L42=L43=39Ω stripline (2.7mm x 3.42mm)
L44=L45=39Ω stripline (8.0mm x 3.42mm)
L46=L49= Ferroxcube wideband hf choke, grade 3B
(4312 020 36642)
L50=L51=50Ω stripline (24.0mm x 3.42mm)

H1=H2= 52mm Sage Wire line semi rigid type BHC

D1= BAW62 or 1N4148

IC1= Voltage stabilizer type μA 7812 UG or smaller version

T1=	BLU98	common-emitter	SOT103 E1
T2=	BLV99	common-emitter	SOT172
T3=	BLV98	common-base	SOT171
T4=T5=	BLV97	common-base	SOT171

All striplines are on a double Cu-clad printed circuit board with PTFE fibre-glass dielectric RT-Duroid type 5880 (ε_r=2.20); thickness 1/32 inch (0.794mm); Cu sheet thickness 2x17μ.

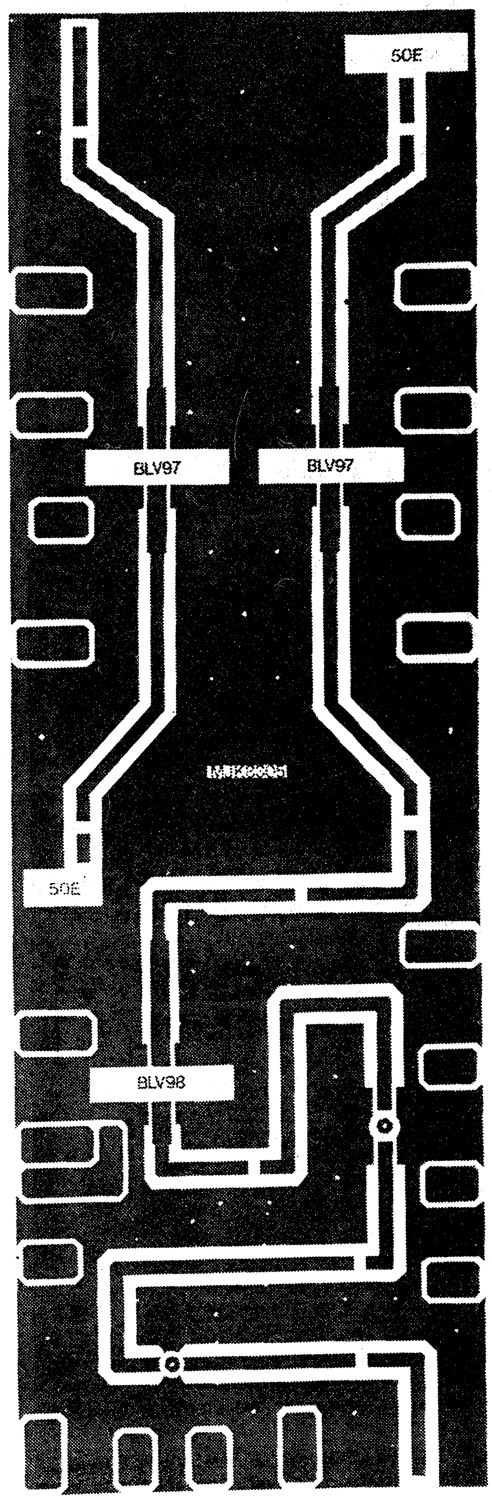


FIG.3

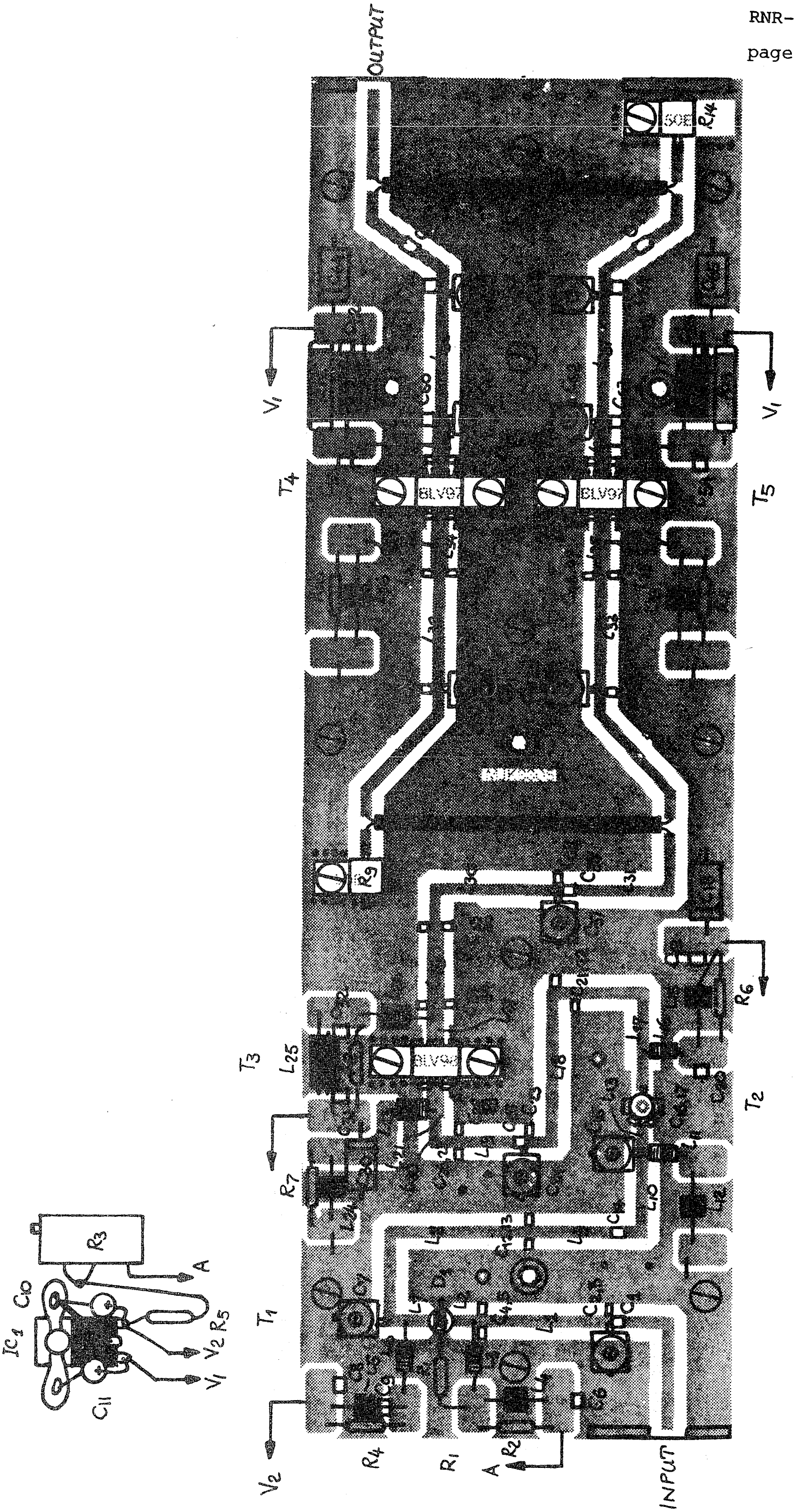

OUTPUT
R14
V1
V1
T4
T5
BLV97
BLV97
R9
R6
T3
L25
T2
R7
T1
R3
A
C10
R5
IC1
V2
V1
C11
V2
R4
R1
A
R2
INPUT

FIG.4

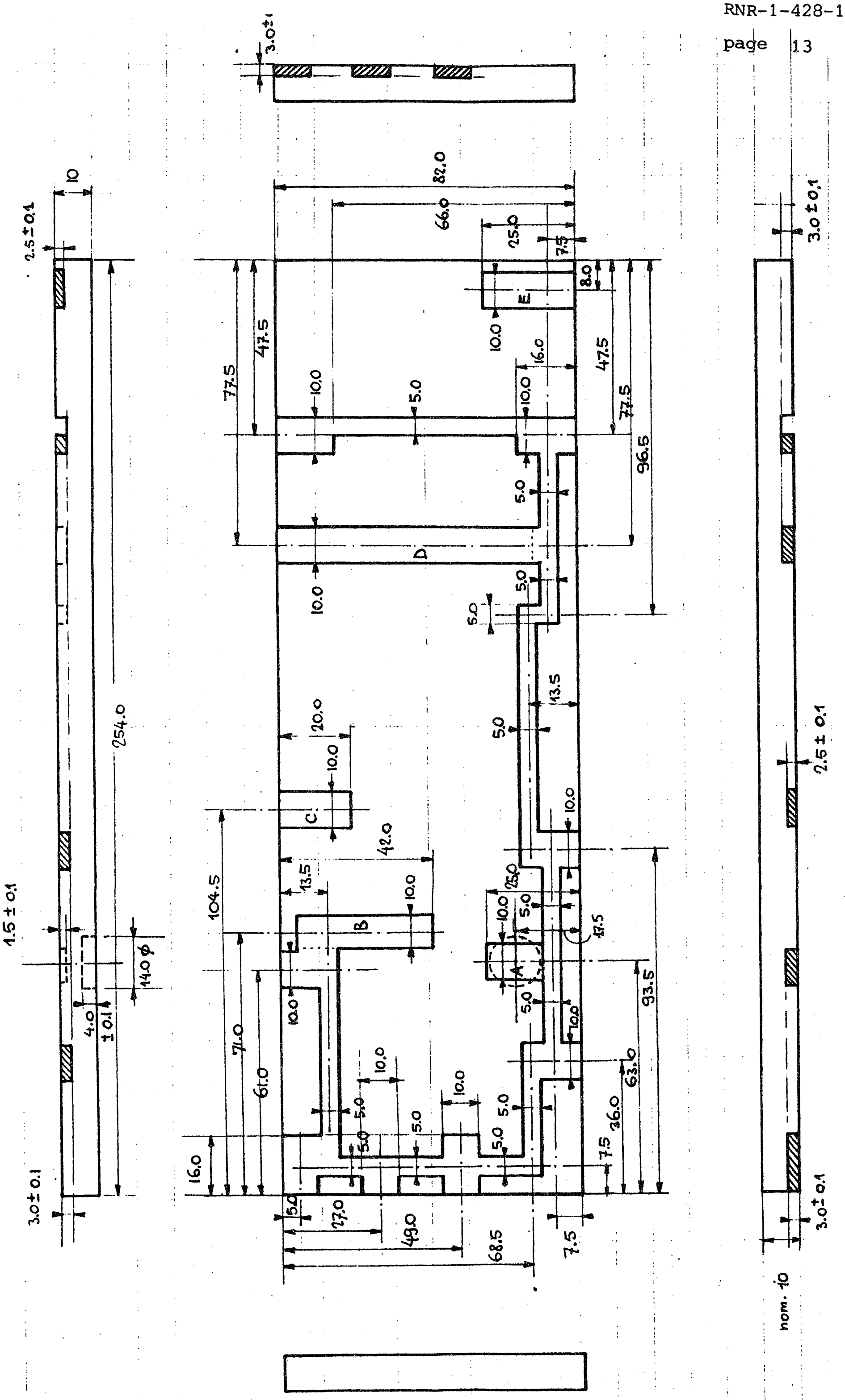


FIG.5

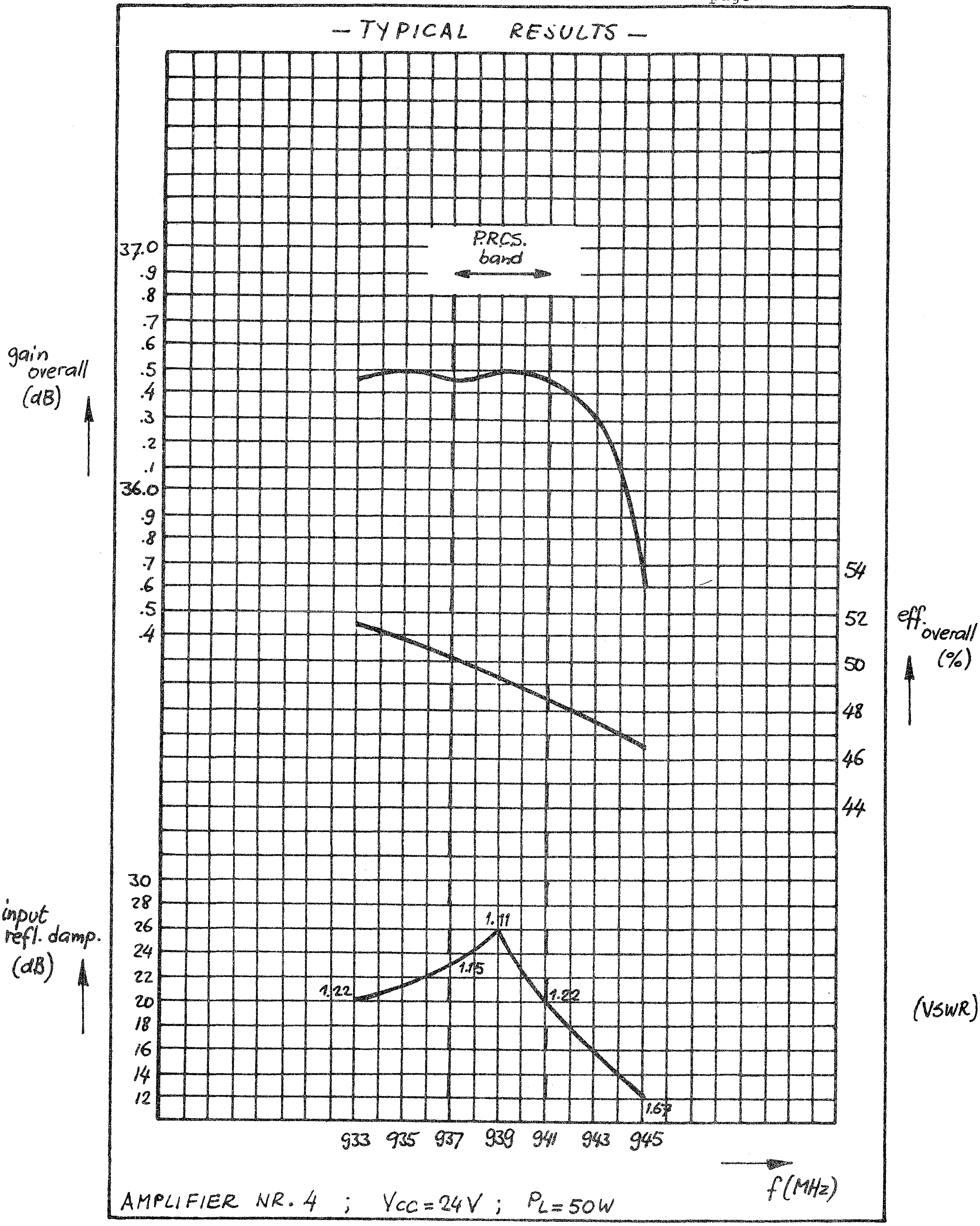

– TYPICAL RESULTS –
P.R.C.S. band
37.0
.9
.8
.7
.6
.5
.4
.3
.2
.1
36.0
.9
.8
.7
.6
.5
.4
gain overall (dB)
54
52
50
48
46
44
eff. overall (%)
30
28
26
24
22
20
18
16
14
12
input refl. damp. (dB)
1.22
1.15
1.11
1.22
1.67
(VSWR)
933
935
937
939
941
943
945
f (MHz)
AMPLIFIER NR. 4 ; Vcc = 24V ; P_L = 50W

FIG.6

Report no : RNR45/008/1988/AS / NCO 8802
Author : R.Gajadharsing
Date : 1988-01-11

400MHz PERFORMANCE OF THE RF POWER MOSFETS BLF242, BLF244 AND BLF245

ABSTRACT

This note provides additional application information about the VHF power MOSFETS BLF242, BLF244 and BLF245. It concerns performance data of these devices at 400MHz. Mean values of powergain and drain efficiency obtained are given in the table below.

	P_o (W)	I_{dq} (mA)	G_p (dB)	Eff. (%)
BLF242	5	10	13.1	59.5
BLF244	15	25	11.3	56.5
BLF245	30	50	10.8	54.0

V_{ds}= 28 Volt; T_h=25°C

Mode of operation: narrowband c.w.

INTRODUCTION

Recently Philips introduced the BLF242, BLF244 and BLF245, the first members of a range of RF-power MOSFETS for professional transmitters capable of delivering output powers of 5W, 15W and 30W respectively. These MOSFETS are primarily designed for operation in FM-transmitters in the VHF-range.
Data sheets of these devices have been published in the 1986 issue of the Philips data handbook "RF power transistors and modules" part S6.
Although these transistors are characterised at 175MHz in these sheets, the high power gains (>13dB) they offer also allow operation in transmitters beyond the VHF-range.
In this note information will be supplied with respect to application at a frequency of 400MHz.

TEST AMPLIFIERS

RF-performance of these transistors has been evaluated in narrowband test amplifiers.
Description of these amplifiers is given on pages 6 to 11.
Calculation of the input and output matching networks has been based on the following impedances:

	BLF242	BLF244	BLF245
Z_i [Ω]	4.9 - j 10.4	2.6 - j 4.5	1.2 - j 2.1
Z_o [Ω]	12.6 - j 26.4	3.3 - j 5.3	1.6 - j 1.0

Z_o is the conjugate of the optimum load impedance into which the device operates at a given output power, voltage and frequency.

The transistor impedances are transformed in two steps to 50Ω, at the input side as well as at the output side. The first step transforms the impedance to 25Ω by means of an LC-section. The second step to 50Ω is obtained by means of a π-section.

In practice minor changes were necessary at the input side to obtain proper matching.
In the BLF242 amplifier L2 had to be increased from 7nH to 9nH, see page 6.

In the BLF244 amplifier the π-section had to be shifted towards the input N-connector in order to increase L3 with L2, see page 8.

For the BLF245 test amplifier an additional coil L2 was needed, see page 10.

Note that resistive input loading has been utilized to enhance stability.

RF-PERFORMANCE

Ten samples of each type, covering the production spread, were used for performance evaluation.
Powergain and drain efficiency have been measured with two different loads presented to the transistors:

1. The optimum load: for each output power the amplifier was tuned for maximum gain.

2. A fixed load: the output network of the amplifier was tuned for a predetermined load and not altered during measurement.

By loading the transistor with a load differing from the optimum one, efficiency improvement has been obtained at the cost of gain. This has not been achieved for the BLF244. The results obtained are given on page 4.
Performance at different output power levels is presented on page 5.
A typical device was chosen for these measurements.

CONCLUSION

For the BLF242, BLF244 and BLF245 it can be stated that at their nominal load power their power gain is greater than 10dB and their drain efficiency greater than 50% at an operating frequency of 400MHz.

R.Gajadharsing

400MHz PERFORMANCE OF THE BLF245 TRANSISTOR

Conditions: Vds= 28V, Po= 30W, Idq= 50mA, Th= 25°C

Sampleno. - Sliceno.	FIXED TUNED LOAD1)					OPTIMALLY TUNED LOAD			
	Ps (W)	Id (A)	Gp (dB)	Eff. (%)		Ps (W)	Id (A)	Gp (dB)	Eff. (%)
13 - 5	3.28	1.64	9.6	65.3		2.52	1.88	10.8	57.0
14	3.42	1.62	9.4	66.1		2.62	1.84	10.6	58.0
16 - 7	2.90	1.78	10.1	60.2		2.56	2.05	10.7	52.3
17	2.75	1.76	10.4	60.9		2.40	2.00	11.0	53.5
24 - 8	2.76	1.75	10.4	61.2		2.26	2.00	11.2	53.5
25	2.64	1.78	10.6	60.2		2.31	2.00	11.1	53.5
28 - 10	2.80	1.83	10.7	58.5		2.43	2.02	10.9	53.0
29	2.70	1.73	10.5	61.9		2.25	1.94	11.2	55.2
35 - 13	3.10	1.81	9.9	59.2		2.76	2.05	10.4	52.3
36	3.12	1.75	9.8	61.2		2.70	2.02	10.5	53.0

1) Load tuned with 13.3 Ohm//80 pF.

400MHz PERFORMANCE OF THE BLF244 TRANSISTOR

Conditions: Vds= 28V, Po= 15W, Idq= 25mA, Th= 25°C

Sampleno. - Sliceno.	OPTIMALLY TUNED LOAD			
	Ps (W)	Id (mA)	Gp (dB)	Eff. (%)
9 - 2	1.23	940	10.9	57.0
10	1.17	955	11.1	56.1
16 - 3	1.06	975	11.4	54.9
17	1.09	966	11.4	55.5
26 - 10	1.02	950	11.7	56.4
27	0.98	955	11.8	56.1
35 - 12	1.11	960	11.3	55.8
36	1.41	905	10.3	59.2
43 - 19	1.01	925	11.7	57.9
44	1.01	925	11.7	57.9

400MHz PERFORMANCE OF THE BLF242 TRANSISTOR

Conditions: Vds= 28V, Po= 5W, Idq= 10mA, Th= 25°C

Sampleno. - Sliceno.	FIXED TUNED LOAD1)					OPTIMALLY TUNED LOAD			
	Ps (mW)	Id (mA)	Gp (dB)	Eff. (%)		Ps (mW)	Id (mA)	Gp (dB)	Eff. (%)
31 - 4	340	282	11.7	63.3		256	293	12.9	60.9
32	345	276	11.6	64.7		250	293	13.0	60.9
33	345	280	11.6	63.8		257	293	12.9	60.9
37 - 10	330	277	11.8	64.5		236	309	13.3	57.8
35 - 18	315	275	12.0	64.9		231	305	13.4	58.5
36	325	278	11.9	64.2		242	300	13.1	59.5
37	320	280	11.9	63.8		241	308	13.2	58.0
36 - 20	385	278	11.4	64.2		260	300	12.8	59.5
37	385	279	11.4	64.0		257	300	12.9	59.5
38	350	280	11.5	63.8		235	302	13.3	59.1

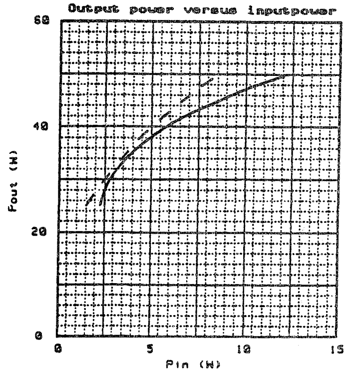
1) Load tuned with 82 Ohm//11.1pF.

BLF245

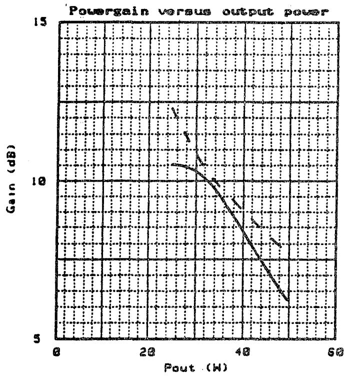
Vds=28V; Idq=50mA; Th=25 °C

—— Fixed tuned load
— — Optimally tuned load

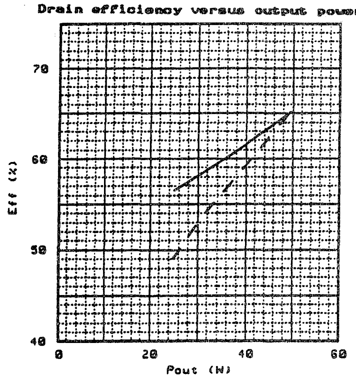
Output power versus inputpower



Powergain versus output power



Drain efficiency versus output power

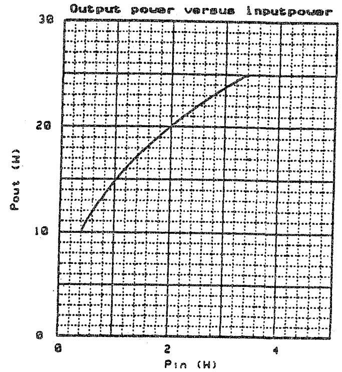


BLF244

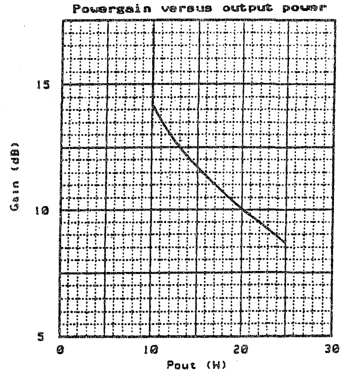
Vds=28V; Idq=25mA; Th=25 °C

—— Optimally tuned load

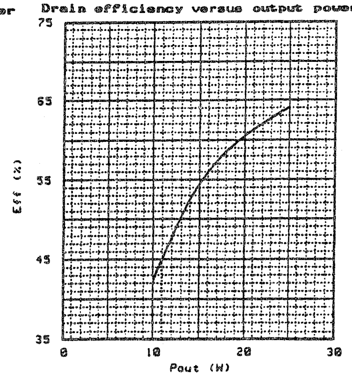
Output power versus inputpower



Powergain versus output power



Drain efficiency versus output power

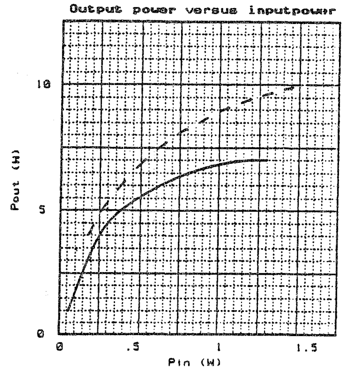


BLF242

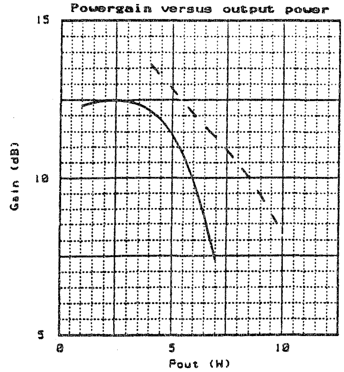
Vds=28V; Idq=10mA; Th=25 °C

—— Fixed tuned load
— — Optimally tuned load

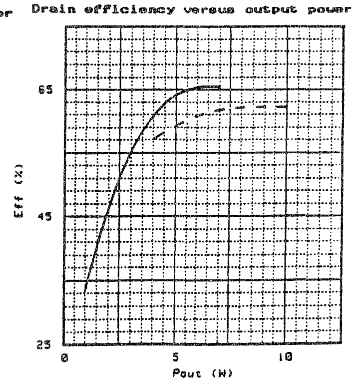
Output power versus inputpower



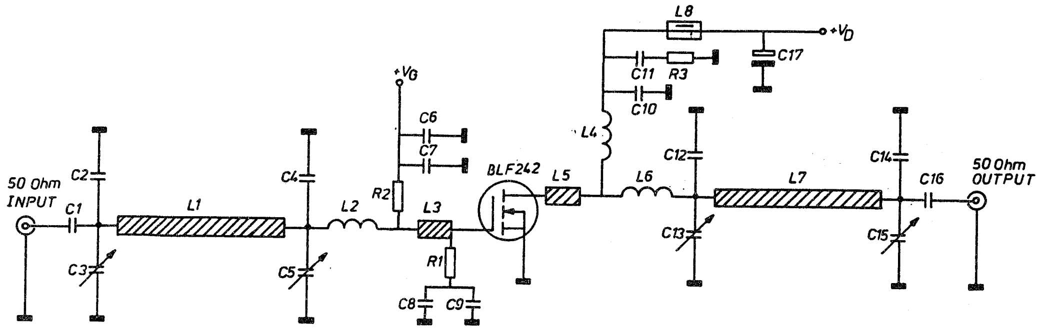
Powergain versus output power



Drain efficiency versus output power



BLF242 400MHZ TEST CIRCUIT



LIST OF COMPONENTS (BLF242)

Capacitors

C1= C7= C8= C10= C16= 560pF	; ceramic multilayer chip capacitor*
C2= C14= 9.1pF	; ceramic multilayer chip capacitor*
C3= C5= C13= C15= 2-18pF	; film dielectric trimmer (cat.no. 2222 809 09003)
C4= 27pF	; ceramic multilayer chip capacitor*
C6= C9= C11= 100nF	; ceramic multilayer chip capacitor(cat.no.2222 852 47104)
C12= 18pF	; ceramic multilayer chip capacitor*
C17= 2.2uF	; Tantalum electrolytic capacitor(35V)

Inductors

L1= L7= 50 Ohm stripline	; (45.7* 4.7)mm
L2= 9nH	; 2 turns enamelled Cu-wire(0.7mm) int.dia= 2mm; l= 2.5mm; leads 2*1mm
L3= 34 Ohm stripline	; (6.0*8.0)mm
L4= 156nH	; 6 turns enamelled Cu-wire(1mm) int.dia= 6mm; l= 8.3mm; leads 2*5mm
L5= 42.5 Ohm stripline	; (5.0*6.0)mm
L6= 13.9nH	; 2 turns enamelled Cu-wire(1mm) int.dia= 3mm; l= 3mm; leads 2*2mm
L8= wideband choke	; ferroxcube grade 3B (cat.no. 4312 020 36642)

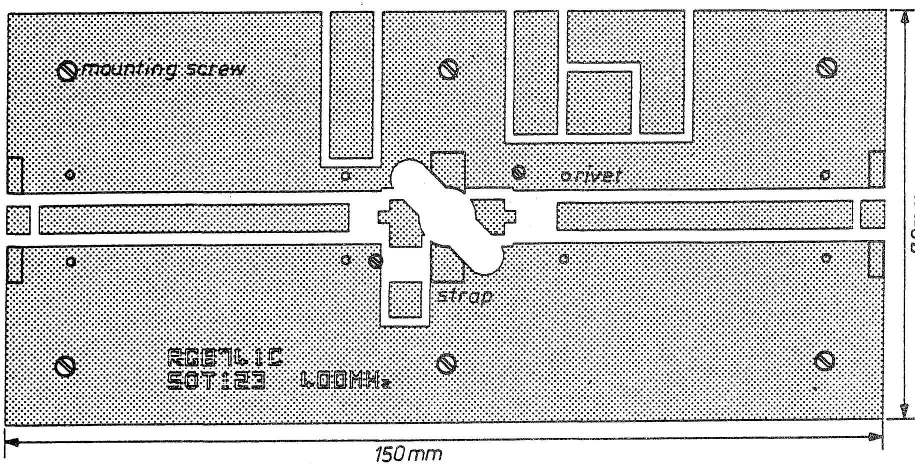
Resistors

R1= 2*133 Ohm in parallel	; metal film resistor(0.4W)
R2= 100 kOhm	; metal film resistor(0.4W)
R3= 10 Ohm	; metal film resistor(0.4W)

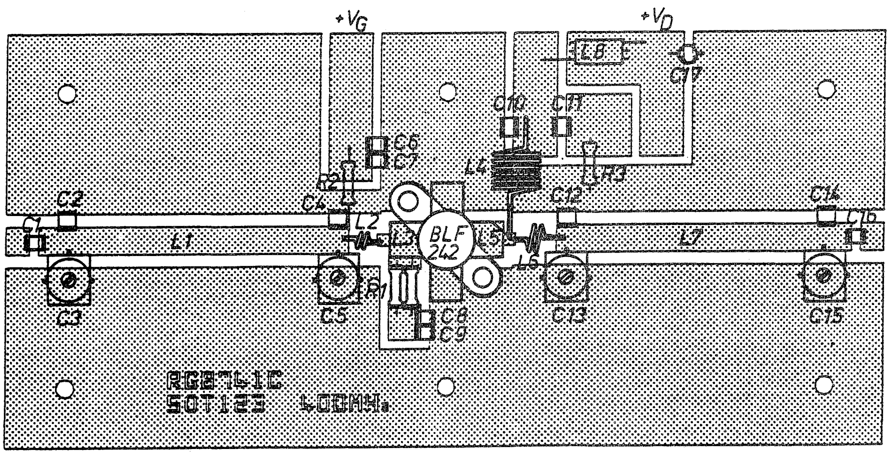
Striplines are on double Cu-clad printed circuit board with glass microfiber reinforced P.T.F.E. dielectric (ε_r= 2.2), thickness 1/16 inch.

*) American technical ceramics capacitor type 100B.

PRINTED CIRCUIT BOARD LAYOUT - BLF242 TEST CIRCUIT

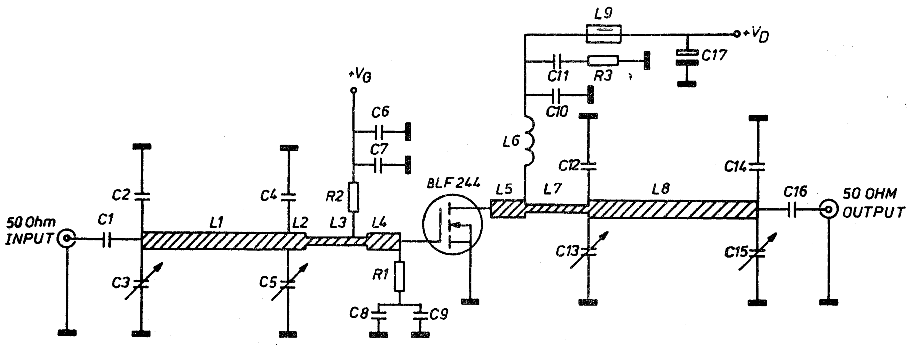


COMPONENT LAYOUT - BLF242 TEST CIRCUIT



THE OTHER SIDE OF THE BOARD IS FULLY METALLIZED AND USED AS GROUND PLANE. THE GROUND PLANES ON EACH SIDE OF THE BOARD ARE CONNECTED TOGETHER BY MEANS OF HOLLOW RIVETS, COPPER STRAPS UNDER THE SOURCE LEADS AND AT THE INPUT AND OUTPUT AND THE MOUNTING SCREWS.

BLF244 400MHZ TEST CIRCUIT



LIST OF COMPONENTS (BLF244)

Capacitors

C1= C7= C8= C10= C16= 560pF	; ceramic multilayer chip capacitor*
C2= 18pF	; ceramic multilayer chip capacitor
C3= C5= C13= C15= 2-18pF	; film dielectric trimmer (cat. no. 2222 809 09003)
C4= 27pF	; ceramic multilayer chip capacitor*
C6= C9= C11= 100nF	; ceramic multilayer chip capacitor(cat. no. 2222 852 47104)
C12= 33pF	; ceramic multilayer chip capacitor*
C14= 9pF	; ceramic multilayer chip capacitor*
C17= 2.2uF	; Tantalum electrolytic capacitor(35V)

Inductors

L1= 50 Ohm stripline	; (44.5* 4.7)mm
L2= 50 Ohm stripline	; (3.5*4.7)mm
L3= 83 Ohm stripline	; (10.8*2.0)mm
L4= 34.5 Ohm stripline	; (5.0*8.0)mm
L5= 42.5 Ohm stripline	; (5.0*6.0)mm
L6= 104nH	; 6 turns enamelled Cu-wire(1mm) int.dia= 4mm; l= 6.6mm; leads 2*5mm
L7= 83 Ohm stripline	; (12.2*2.0)mm
L8= 50 Ohm stripline	; (45.7*4.7)mm
L9= wideband choke	; ferroxcube grade 3B (cat. no. 4312 020 36642)

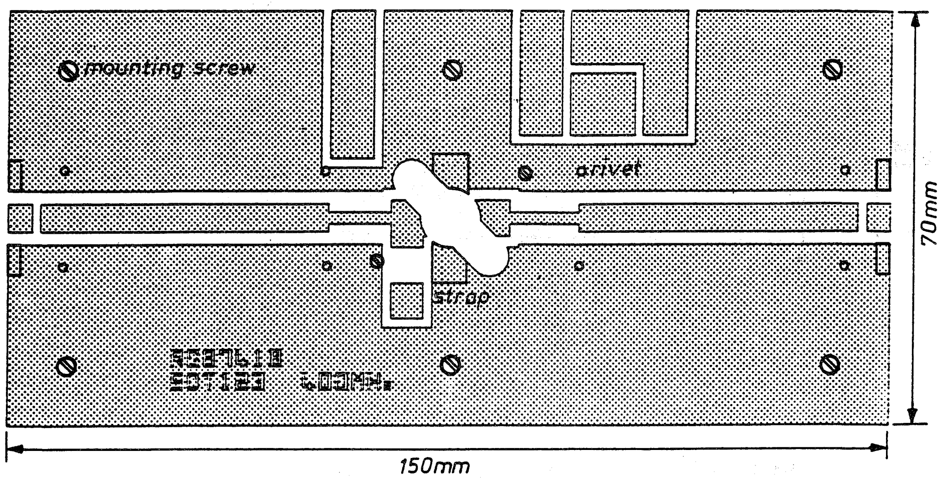
Resistors

R1= 2*53.6 Ohm in parallel	; metal film resistor(0.4W)
R2= 100 kOhm	; metal film resistor(0.4W)
R3= 10 Ohm	; metal film resistor(0.4W)

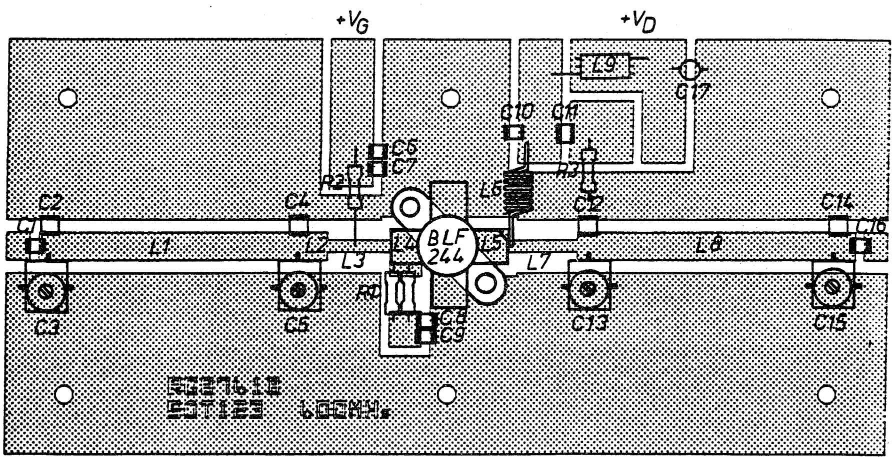
Striplines are on double Cu-clad printed circuit board with glass microfiber reinforced P.T.F.E. dielectric (ε_r= 2.2), thickness 1/16 inch.

*) American technical ceramics capacitor type 100B.

PRINTED CIRCUIT BOARD LAYOUT - BLF244 TEST CIRCUIT



COMPONENT LAYOUT - BLF244 TEST CIRCUIT



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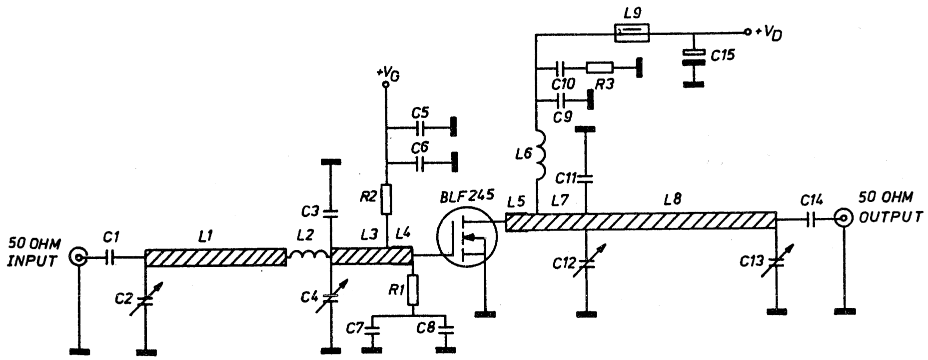
BLF245 400MHZ TEST CIRCUIT



LIST OF COMPONENTS (BLF245)

Capacitors

C1= C6= C7= C9= C14= 560pF	; ceramic multilayer chip capacitor*
C2= 2-9pF	; film dielectric trimmer (cat. no. 2222 809 09002)
C3= 27pF	; ceramic multilayer chip capacitor*
C4= C12= C13= 2-18pF	; film dielectric trimmer (cat.no. 2222 809 09003)
C5= C8= C10= 100nF	; ceramic multilayer chip capacitor(cat.no.2222 852 47104)
C11= 43pF	; ceramic multilayer chip capacitor*
C15= 2.2uF	; Tantalum electrolytic capacitor(35V)

Inductors

L1= 50 Ohm stripline	; (31.0* 4.7)mm
L2= 19.9nH	; 3 turns enamelled Cu-wire(1mm) int.dia= 3mm; l= 6mm; leads 2*3mm
L3= 50 Ohm stripline	; (9.5*4.7)mm
L4= 34.5 Ohm stripline	; (5.0*8.0)mm
L5= 42.5 Ohm stripline	; (5.0*6.0)mm
L6= 52nH	; 4 turns enamelled Cu-wire(1mm) int.dia= 4mm; l= 6mm; leads 2*5mm
L7= 50 Ohm stripline	; (8.2*4.7)mm
L8= 50 Ohm stripline	; (45.7*4.7)mm
L9= wideband choke	; ferroxcube grade 3B (cat.no. 4312 020 36642)

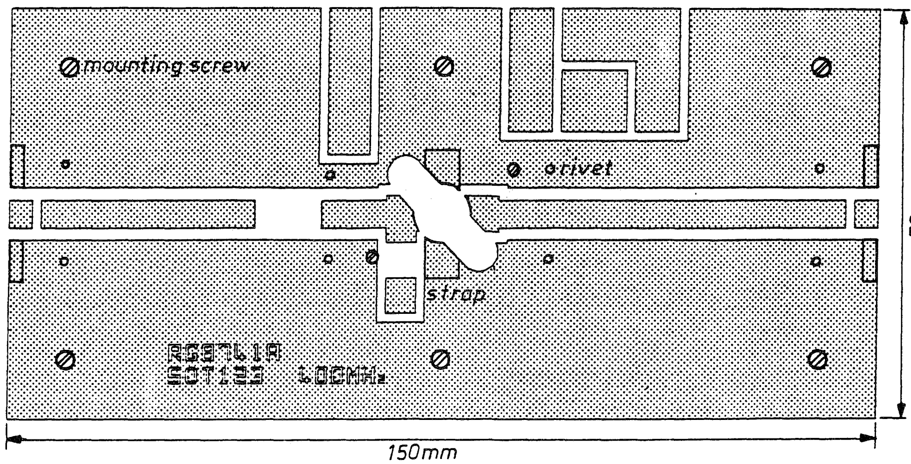
Resistors

R1= 100 Ohm	; metal film resistor(0.4W)
R2= 100 kOhm	; metal film resistor(0.4W)
R3= 10 Ohm	; metal film resistor(0.4W)

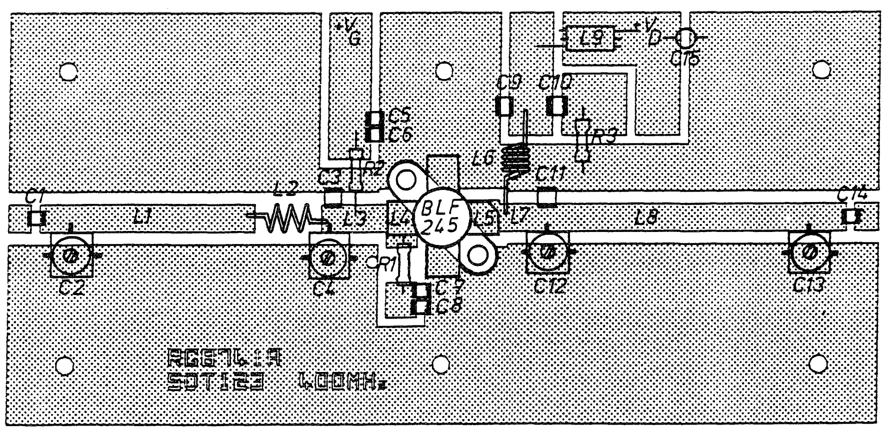
Striplines are on double Cu-clad printed circuit board with glass microfiber reinforced P.T.F.E. dielectric (ε_r= 2.2), thickness 1/16 inch.

*) American technical ceramics capacitor type 100B.

PRINTED CIRCUIT BOARD LAYOUT - BLF245 TEST CIRCUIT



COMPONENT LAYOUT - BLF245 TEST CIRCUIT



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ECO 6907

application information

530

Design of H.F. Wideband Power Transformers

A. H. Hilbers

Date of release: 17 June 1970

An analysis is given of the wideband transmission line transformer with special emphasis on the power handling capability.
Compensation techniques for extending the frequency range in both directions are discussed.
Some practical examples are given of transformers for wideband s.s.b. transmitters in the frequency range 1.6 to 28 MHz, having power handling capabilities up to 80 W.

1 INTRODUCTION

Transmission line power transformers can be used to perform a variety of functions, among which are phase reversal, balanced to unbalanced coupling, impedance transformation and hybrid functions. Such transformers find many applications in wide-band power amplifiers for both s.s.b. transmitters in the h.f. region and f.m. transmitters in the lower v.h.f. region.

The properties of a practical h.f. power transformer are discussed here and their effect on transformer performance is analysed. Since losses must be kept low, in practice the transformer will use a ferrite core. Further, we have limited the discussion to cores without an air-gap since these have a low stray magnetic field, a high permeability, and can cover the power range (up to 80 W) dealt with here. Data (dimensions, permeability values etc.) on all core types can be found in our Data Handbook "Components and Materials", Part 4. A glance through the Handbook will show the wide range of materials, dimensions and types from which the designer may choose. It must be remembered, of course, that when cores constructed in two parts (pot-cores and cross-cores, for example) are used, the type without an air-gap must be selected.

Throughout we have aimed at giving practical solutions to the problems posed by material and design limitations. In particular, compensating techniques for extending the frequency range of a number of transformer configurations are discussed. To give an idea of some application possibilities, practical examples in several transformer configurations have been worked, using transformer cores from our range of ferrites.

2 TRANSFORMER SPECIFICATION

The transformer design considerations dealt with in this publication are:

- maximum power level to be handled
- frequency range
- input and output impedance
- allowable reflection and resistive losses.

How a transformer can meet the above considerations for a particular application is analysed in the following three sections. The first two sections deal with the influence of the core and transmission line respectively on transformer performance, and the third with mismatch compensation techniques.

3 INFLUENCE OF THE CORE ON PERFORMANCE

3.1 Primary Inductance

This inductance determines the amount of reflection at the low frequency end of the band. It can be calculated using the formula:

$$L = \mu_o \mu_r n^2 A/l \tag{1}$$

in which L = inductance in H
$\mu_o = 4\pi\ 10^{-7}$ (rationalised M.K.S. units.)
μ_r = relative permeability
A = average ferrite cross section in m^2
l = average length of the lines of force in m.
n = number of turns between the input connections.

In a simple example, like the phase reversing transformer, this relation holds. Other cases may require a transformation (see Sub-Section 7.1).

If degrading of performance at the high end of the band is to be avoided, the value of L must not be higher than really necessary. A good practical value is:

$$L = 4R/\omega_{min} \tag{2}$$

in which R = midband input resistance in Ω
ω_{min} = 2π times the minimum frequency in Hz.

Where requirements are severe the compensation technique described in Sub-Section 5.1 may be used.

3.2 Core Losses

The losses caused by the core material will be represented here as a resistance (R_p) in parallel with the input. This resistance depends on:

- the sort of ferrite material
- the frequency
- the quantity L/μ_r
- the maximum flux density B_{max}.

In the small signal case ($B_{max} \to 0$), R_p can be calculated with the aid of curves of the type shown in Fig. 1*. In these curves a comparison is made between different core materials based on equal core dimensions and equal number of turns. It can be seen that 4C4 and 4C6 are the best materials for frequencies above approximately 2.5 MHz. In the high v.h.f. region 1Z2 ferroxplana shows interesting properties as can be seen from the same figure.

The power handling capability of a transformer is closely dependent on the behaviour of R_p as a function of B_{max}. For the section of the B-H curve with which we are dealing, B_{max} can be calculated using the formula:

$$B_{max} = V_{max}/\omega \cdot A \cdot n \tag{3}$$

in which B_{max} = maximum flux density in T**
ω = 2π times frequency in Hz
A = ferrite cross section in m^2
n = number of turns
V_{max} = maximum value of voltage across n turns in V.

---

* The curves of Figs 1 and 2 have been drawn from measurements on single samples of the ferrite materials. Thus the average curves may differ somewhat from those shown.

** The letter T stands for Tesla, the unit of magnetic flux density in the SI unit system. The following relationship holds:

$$1\ T = 1\ Wb/m^2 = 1\ Vsec/m^2 = 10\ 000\ gauss.$$

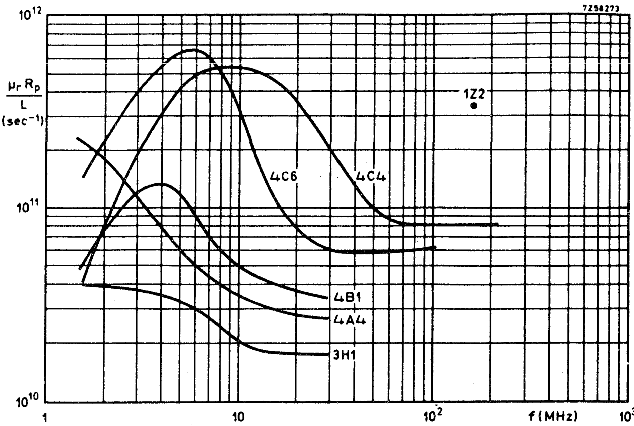


Fig.1. Curves of $\mu_r R_p/L$ plotted against frequency for our various ferrites. The curves have been plotted for small signal conditions ($B_{max} \to 0$).

In Fig.2, the quantity $\mu_r R_p/L$ is given for different ferrite materials as a function of the product $B_{max} \cdot f$ with the frequency as a parameter. The product $B_{max} \cdot f$ has been chosen because, for most transformers, its value remains constant for changing frequency. From Fig.2 it can be seen that R_p decreases as B_{max} increases, especially at lower frequencies. This forms the primary limit on the power handling capability of these transformers. If 4C4 material (Fig.2*d*) is used in the h.f. region, the $B_{max} \cdot f$ product must not be higher than approx. 2 x 10^4 T.Hz. Combining this with the choice of L according to eq.(2), we find that the power loss caused by the core material will be no more than 1 %. At frequencies of 30 MHz and higher it seems that higher $B_{max} \cdot f$ products, perhaps up to 10^5 T. Hz, can be used. For 1Z2 ferroxplana this has already been confirmed by measurements at 165 MHz.

A very conservative choice of the B_{max} value must also be avoided because this leads to a greater length of the transmission line and consequently more loss at the high end of the band.

4 INFLUENCE OF THE TRANSMISSION LINE ON PERFORMANCE

4.1 Resistive Loss and Power Handling

The power loss in the transmission line depends on:

- the type of line
- the frequency
- the length.

Data on power loss in some 50 Ω coaxial cables is given in Fig.3 (p.10). This power loss and the allowable maximum cable temperature restrict the power handling of the cable.
The maximum power which can be transmitted depends on the type of cable and the frequency; data is given in Fig.4 (p.10).

4.2 Mismatch Loss

Another kind of loss caused by the transmission line can occur if the characteristic impedance of this line is not the required value. This results in a mismatch being maximum at the high frequency end of the band. The amount of mismatch depends on:

- the ratio between the length of the line and the wavelength on the line
- the ratio between the required and the actual value of the characteristic resistance.

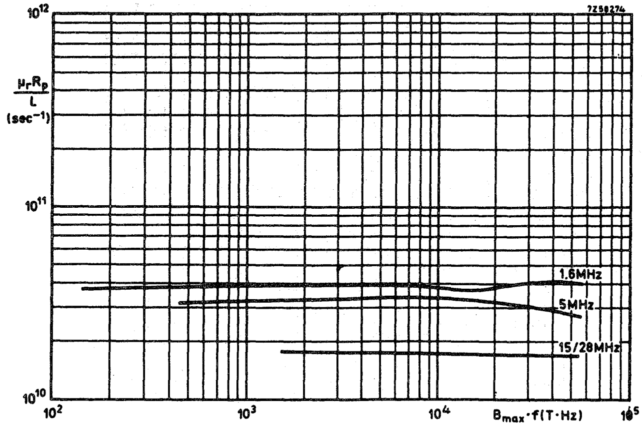


Fig.2*a*. 3H1 material.

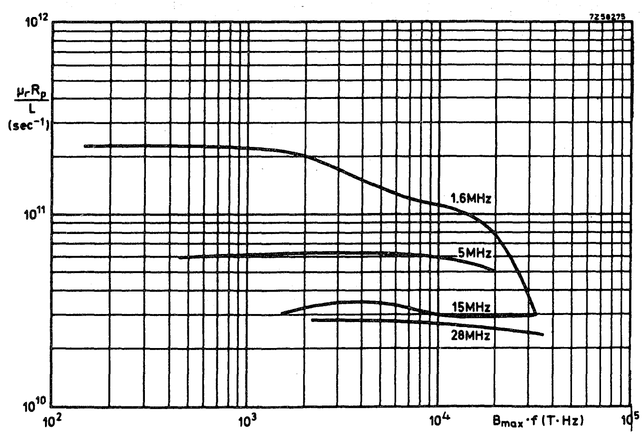


Fig.2*b*. 4A4 material.

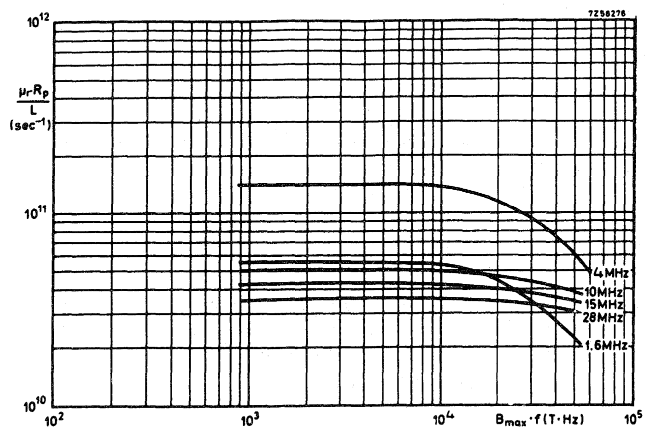


Fig.2*c*. 4B1 material.

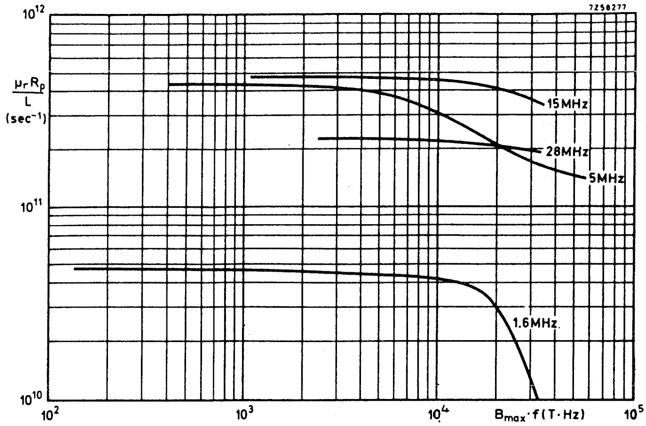


Fig. 2*d*. 4C4 material.

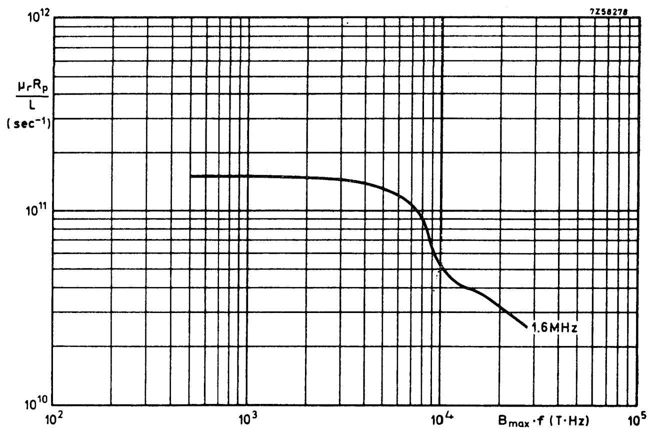


Fig. 2*e*. 4C6 material.

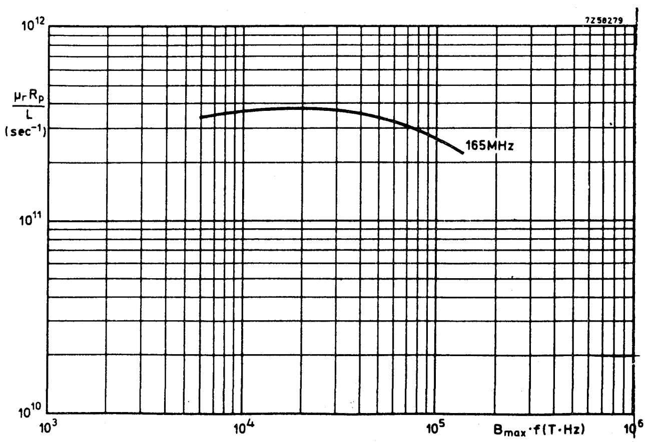


Fig. 2*f*. 1Z2 material.

Fig. 2. Curves of $\mu_r R_p / L$ plotted against $B_{max} \cdot f$ with frequency as parameter for our various types of ferrite.

9

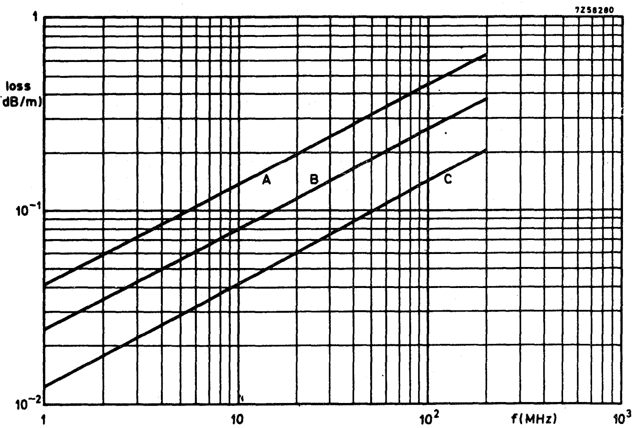


Fig.3. Curves of power loss plotted against frequency for three 50 Ω coaxial cables.
A: diameter = 1.7 mm. B: diameter = 2.8 mm. C: diameter = 5 mm.

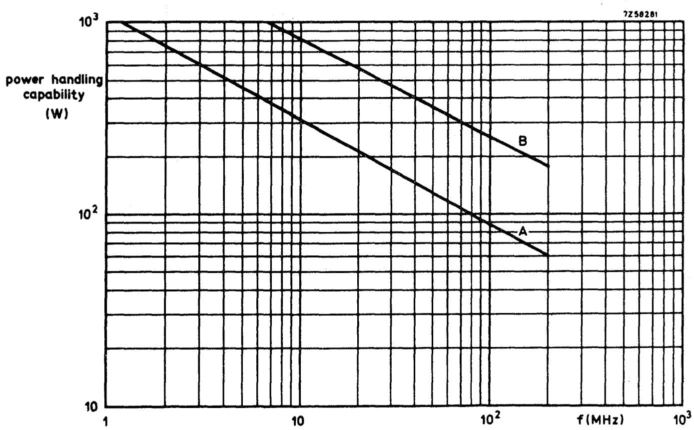


Fig.4. Curves of power handling capability plotted against frequency for two 50 Ω coaxial cables.
A: diameter = 2.8 mm. B: diameter = 5 mm.

From transmission line theory the input impedance is given by:

$$Z_{in} = R_{in} \cdot \frac{1 + j\, r \tan \beta l}{1 + j \frac{1}{r} \tan \beta l} \tag{4}$$

in which R_{in} = midband input resistance
r = ratio between the actual and the required characteristic resistance
β = $2\pi/\lambda$
λ = wavelength on the line (for 50 Ω coaxial cables approx. 67 – 70 % of wavelength in free space)
l = length of the line.

10 If the deviation of Z_{in} from the required value is unacceptably large it is in many cases possible to make use of the compensation technique described in Sub-Section 5.2.

5 COMPENSATION TECHNIQUES

5.1 Compensation At Low Frequencies

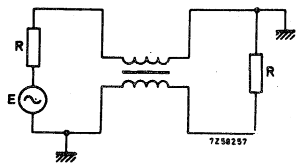


Fig.5*a*. Schematic diagram.

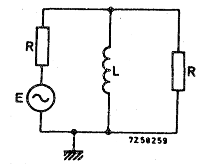


Fig.5*b*. Low frequency equivalent diagram.

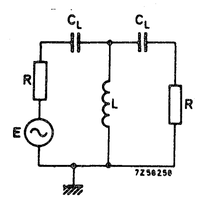


Fig.5*c*. Equivalent diagram with compensation.

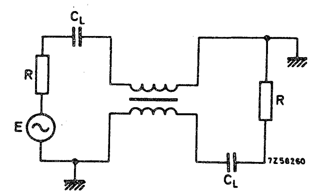


Fig.5*d*. Schematic diagram with compensation.

Compensation will be illustrated by means of the phase reversing transformer. The schematic diagram is given in Fig.5*a* and the equivalent diagram for the low frequency end of the band is shown in Fig.5*b*. For compensation we add two equal capacitors C_L such that a high-pass T-filter section is formed (Fig.5*c*). According to filter theory:

$$C_L = 2L/R^2 \tag{5}$$

The original diagram of Fig.5*a* is now transformed to the new one of Fig.5*d*. If L is dimensioned according to eq(2) the input impedance without compensation at the lowest frequency is $R//(+j\,4R)$. With compensation the input impedance is $(.999R// +j\,264\,R)$, illustrating that the mismatch has been reduced to a negligible level.

For some types of transformer, the capacitor at the output must have a different value from that at the input. With a $1 : n^2$ impedance transformer, for instance, it must be n^2 times smaller than the capacitor at the input.

Sometimes two or more transformers must be connected in cascade. In such a case low frequency compensation is possible if a high-pass π-filter section is used (Fig.5*e*). When the parallel inductances of the transformers at the interconnecting point are both approximately equal to L, the capacitance of C_L must be $L/2R^2$.

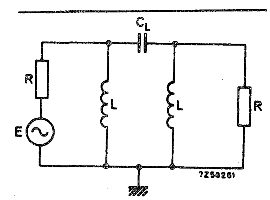


Fig.5*e*. Equivalent diagram of cascaded transformers with compensation.

Fig.5. Low frequency compensation of the phase reversing transformer.

5.2 Compensation at High Frequencies

This is only necessary when the characteristic resistance of the transmission line differs from the required value. A situation often met with in practice is that in which the required characteristic resistance is lower than that of the available line. Taking the simple case of the phase reversing transformer (Fig.6) with a required characteristic resistance equal to R, we find that compensation for an actual value equal to $r \times R$ can be made as follows. In parallel with the load resistance, we connect a capacitor of such a value that at the highest frequency the real part of the input admittance becomes $1/R$, The resulting imaginary part of the input admittance is tuned out by means of a capacitor in parallel with the input. Both capacitors turn out to have the same value, given by

$$C_H = \frac{1}{\omega_{max}\, rR \tan \beta l} \left\{ 1 - \sqrt{1 - (r^2 - 1) \tan^2 \beta l} \right\} \tag{6}$$

in which $\omega_{max} = 2\pi$ times the maximum frequency. The schematic diagram is shown in Fig.6.

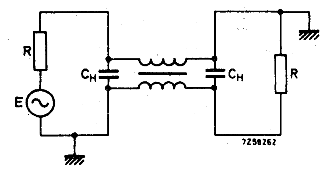


Fig.6. Phase reversing transformer with high frequency compensation.

The result of this compensation is an exact match at the maximum frequency. There will be however, a slight mismatch at lower frequencies which is many times smaller than that at the maximum frequency without compensation. A combination of high and low frequency compensation is of course possible.

6 TRANSFORMER CONFIGURATION

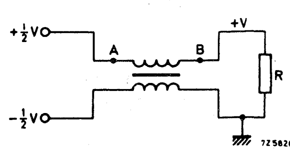
Because of the variety of the existing configurations it is hardly possible to give a complete survey. Therefore a restriction will be made to some principally different types.

6.1 Phase Reversing Transformer

This type has already been discussed in the previous section.

6.2 Balanced to Unbalanced Transformer

The schematic diagram is shown in Fig.7. This type can be considered as a modification of the phase reversing transformer. The primary inductance in this case is 4 times the inductance of the winding between the points A and B because of the voltage division. If low frequency compensation is used, a capacitor equal to $2C_L$ must be placed in series with each input connection (see Section 5.1) to preserve symmetry, and one capacitor equal to C_L in series with the output (point B).



12

Fig.7. Balanced to unbalanced transformer.

6.3 Symmetrical* 1 : 4 Impedance Transformer

The schematic diagram is given in Fig.8. Two cables each having a characteristic resistance of $2R$, can be wound on a common core. The direction of the windings follows from the voltage division. Low frequency compensation can be made with a capacitor equal to $2C_L$ (see Section 5.1) in each of the input leads and a capacitor equal to $1/2C_L$ in each of the output leads.

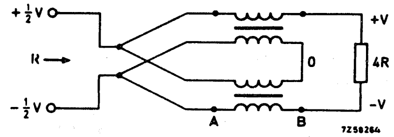


Fig.8. Symmetrical 1 : 4 impedance transformer.

6.4 Asymmetrical 1 : 4 Impedance Transformer

If, in the 1 : 4 impedance transformer of section 6.3, the points A and B are connected to earth it is no longer possible to wind the two lines on a common core. In fact the lower line may be wound on an "air core" because there is no voltage difference between the points A and B. The logical next step is to omit the lower line completely (with a consequent slight phase difference between the lines). Then we get the transformer shown in Fig.9. The characteristic resistance of the transmission line has again an optimum value of $2R$. But even if this value is chosen the input impedance is not constant as a function of frequency. From theory (see ref.5) it is:

$$Z_{in} = R \times \frac{2 \cos \beta l + jr \sin \beta l}{1 + \cos \beta l + j\frac{1}{r} \sin \beta l} \qquad (7)$$

in which r = ratio between the actual characteristic resistance and $2R$.

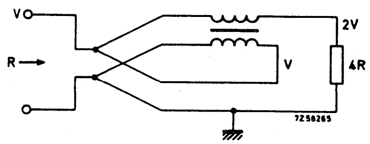


Fig.9. Asymmetrical 1 : 4 impedance transformer

If $r > 1$ high frequency compensation is sometimes possible. A capacitor C_1 given by:

$$C_1 = \frac{1 + \cos \beta l - \sqrt{(1 + \cos \beta l)^2 - r^2 \sin^2 \beta l}}{\omega_{max}\, rR \sin \beta l} \qquad (8)$$

must then be connected across the input and a capacitor C_2 given by:

$$C_2 = \frac{2 \cos \beta l - \sqrt{(1 + \cos \beta l)^2 - r^2 \sin^2 \beta l}}{4\omega_{max}\, rR \sin \beta l} \qquad (9)$$

must be connected across the output.

* Terminology in normal use is employed to describe the various transformer configurations. However the terms "symmetrical", "balanced" and "push-pull" are used here synonymously to mean "antiphase port signals having equal amplitudes with respect to ground", while likewise "asymmetrical", "unbalanced" and "single-ended" are synonymous with "one port terminal grounded".
Strictly defined, the terms "asymmetrical" and "unbalanced" also apply to unequal port terminal values with respect to ground.

6.5 Symmetrical 9 : 1 Impedance Transformer

The schematic diagram is given in Fig.10. The two transmission lines have an optimum characteristic resistance of $3R$. They can again be wound on a common core, and the third line omitted as in the previous case. The input impedance is given by:

$$Z_{in} = 9R \times \frac{4 + 5\cos\beta l + j\,6\,r\sin\beta l}{9\cos\beta l + j\frac{6}{r}\sin\beta l} \tag{10}$$

in which r is the ratio between the actual characteristic resistance and $3R$.
If $r > 1$ high frequency compensation is sometimes possible. A capacitor C_1 given by:

$$C_1 = \frac{4 + 5\cos\beta l - \sqrt{9(3\cos^2\beta l + 4\cos\beta l + 2) - 36\,r^2\sin^2\beta l}}{6\omega_{max}\,rR\sin\beta l} \tag{11}$$

must than be connected across the low impedance side, and a capacitor C_2, given by:

$$C_2 = \frac{9\cos\beta l - \sqrt{9(3\cos^2\beta l + 4\cos\beta l + 2) - 36\,r^2\sin^2\beta l}}{54\omega_{max}\,rR\sin\beta l} \tag{12}$$

must be connected across the high impedance side.

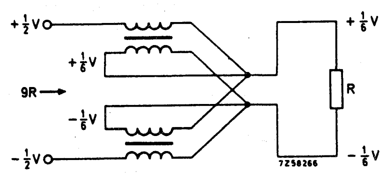


Fig.10. Symmetrical 9 : 1 impedance transformer.

6.6 Asymmetrical 1 : 9 Impedance Transformer

In this transformer (see Fig.11), two transmission lines are required, each having an optimum characteristic resistance of $3R$. Although these lines can be wound on a common core, the upper line must have twice the number of turns of the lower line, because of the voltage division. The third line has again been omitted.

High frequency compensation follows the same principle as that used in previous sections.

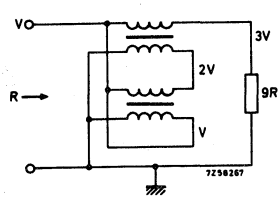


Fig.11. Asymmetrical 1 : 9 impedance transformer.

6.7 Single-ended Hybrid

This circuit (see Fig.12) permits the combination of two signals in a common load ($R/2$) in such a way that the signal sources do not influence each other. When the two signals have different frequencies the power is equally divided between the resistors $R/2$ and $2R$. When the two signals have the same frequency, phase and amplitude, all the power is delivered to

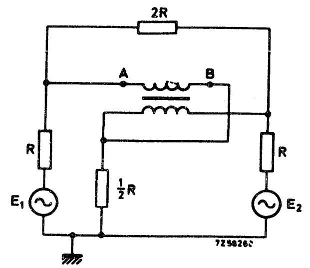


Fig.12. Single-ended hybrid.

$R/2$. The optimum value of the characteristic impedance of the transmission line is equal to R. At the low frequency end of the band the isolation between the sources depends on the inductance of the winding. When the isolation is expressed as a power ratio S the inductance between the points A and B must be:

$$L = \frac{R}{8\omega_{min}} \sqrt{S - 1}\,. \tag{13}$$

Even when the optimum characteristic impedance is chosen, high frequency compensation has been found to have a marked influence on the isolation. Practically it has been found that a small capacitor must be connected in parallel with $R/2$.

To illustrate the degree of isolation that may be expected in a practical design, an example is given of a hybrid transformer designed for the frequency band of 1.6 to 28 MHz. The requirement was to combine two signals of different frequency and equal power (3 W in 100 Ω) in a single 50 Ω load with a minimum isolation of 40 dB between the sources. From eq.(13), it follows that L must be 125 μH minimum. Because of this high inductance it was decided to choose 3H1 material. An available 23 x 14 x 7 mm toroid belonged to "group 5" which meant that the minimum μ_r was 2680.* The required number of turns was 9 The optimum value of the characteristic resistance is 100 Ω, but for reasons of convenience 150 Ω miniature twin lead was used. The calculated value of B_{max} was 87 gauss at 1.6 MHz. High frequency compensation was achieved by means of a 33 pF capacitor in parallel with the 50 Ω load resistance. The isolation between the sources measured as a function of frequency is given in table 1 below, illustrating that the minimum isolation requirement of 40 dB has been fulfilled over the desired frequency band.

Table 1.

f (MHz)	isolation (dB)
0.5	33
1	41
1.6	42.5
5	40.5
15	41
28	43
40	32

6.8 Push-pull Hybrid

Two different versions will be discussed. In the first one the impedance level is stepped-up by a factor of two and in the second the impedance level is stepped-down by a factor of two.

6.8.1 Impedance Step-up Type

The two transmission lines (see Fig.13) have an optimum characteristic resistance equal to R and they can be wound on a common core.

*For the convenience of the user the toroids of ferroxcube 3H1 are delivered sorted into groups of approximately equal μ-value. The μ-value is indicated by the colour of the circumference of the toroids, see Data Handbook System. Groups are not separately available.

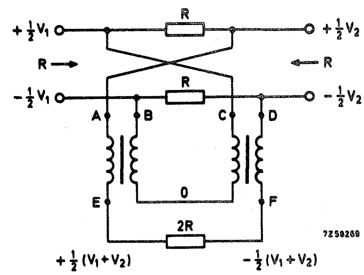


Fig.13. Push-pull hybrid with impedance step-up.

A hybrid can always be used in two ways. The first way has been mentioned in Section 6.7. The second way is the inverse of the first i.e. the power from a single source is equally divided between two loads which are isolated from each other. The latter arrangement allows a more convenient calculation of the lower frequency limit. To do this we must know the effective inductance in parallel with $2R$. The voltage across this resistor is 4 times the voltage between the points D and F and therefore the effective inductance in parallel with $2R$ is 16 times the inductance between the points D and F. For low frequency compensation the effective inductance in parallel with $2R$ can be considered as two equal inductances in series – one between points E and O and one between points F and O – each having a value of 8 times the inductance between points D and F. The hybrid can then be considered as two balanced to unbalanced transformers on a common core and the compensation can be made as described in Sub-Section 6.2.

6.8.2 Impedance Step-Down Type

This hybrid (see Fig.14) is identical to the previous one except that the transmission lines on one side have been connected in parallel instead of in series. When this hybrid is used for the combination of two signals which have the same frequency, phase and amplitude, the ferrite core will not be magnetized. However when V_2 becomes zero, the voltage across each winding will be $V_1/4$ and the voltage across $R/2$ will be $V_1/2$. This means that the effective inductance in parallel with $R/2$ becomes 4 times the inductance of a single winding. A similar approach must be used in finding the winding V/turn when calculating B_{max}.

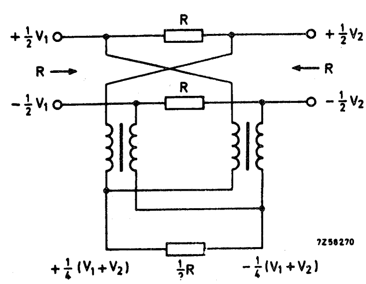


Fig.14. Push-pull hybrid with impedance step-down.

7 PRACTICAL EXAMPLES

It is possible to combine some of the functions, mentioned in Section 6, in one transformer. In the two examples given, simple calculations show the method used in designing for a particular application.

7.1 12.5 Ω Balanced to 50 Ω Unbalanced Transformer

In this case, the transformers described in Sections 6.2 and 6.3 have been combined. The schematic diagram is shown in Fig.15. A transformer of this type is required for a wideband s.s.b. transmitter in the frequency range of 1.6 to 28 MHz. It must be able to handle a peak envelope power of 80 W. The input impedance is 12.5 Ω balanced and the output impedance is 50 Ω unbalanced. The total amount of resistive and reflection losses is required to be below 5 % (power).

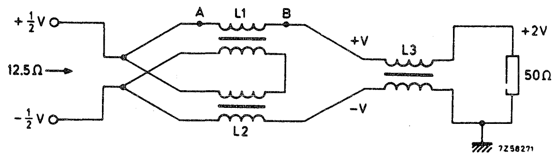


Fig.15. 12.5 Ω balanced to 50 Ω unbalanced transformer.

The transformer has been wound on a single 4C4 toroid of 36 x 23 x 15 mm. Windings L_1 and L_2 must have a characteristic resistance of 25 Ω; they consist of two 50 Ω coaxial cables of 2.8 mm diameter in parallel. Winding L_3 must have a characteristic resistance of 50Ω; a single 50Ω coaxial cable of 2.8 mm diameter has been used. The winding direction follows from the voltage division. From this it is also clear that windings L_1 and L_2 must have an equal number of turns and that winding L_3 must have twice the number of winding L_1.

We can calculate the lower frequency limit from the effective inductance in parallel with the 50 Ω load resistance. We allow a reactance of +j 200 Ω at 1.6 MHz, corresponding to an inductance of 20 μH, in parallel with the 50 Ω load (see Section 3.1). As the voltage between points A and B is ¼ of the output voltage, the inductance between these points will be 20/16 = 1.25 μH. The corresponding number of turns is:

$$n = \sqrt{\frac{L \cdot l}{\mu_0 \mu_r A}}$$

in which l/A = 9.42 cm^{-1} and μ_r is taken to be 100. This gives:

$$n = 3.06.$$

We take n = 3.5 turns, making the inductance in parallel with the 50 Ω load resistance equal to:

$$20\left(\frac{3.5}{3.06}\right)^2 = 26.2\ \mu\text{H minimum.}$$

The measured value was actually 39 μH, because the μ_r of the core used was greater than 100.

The maximum flux density B_{max} can be calculated when the maximum voltage across the windings is known. The peak value of the voltage across the 50 Ω load resistance for an 80 W power level is:

$$V_p = \sqrt{2 \times 80 \times 50} = 89.5\ \text{V},$$

and between points A and B: 89.5/4 = 22.35 V. The ferrite cross section for this core is 0.976 x 10^{-4} m^2 So at 1.6 MHz the maximum flux density is:

$$B_{max} = \frac{V_{max}}{\omega A n} = \frac{22.35}{2 \times \pi \times 1.6 \times 10^6 \times 0.976 \times 10^{-4} \times 3.5} = 65.4 \times 10^{-4}\ \text{T}$$ 17

or 65.4 gauss.

This corresponds to a $B_{max} \cdot f$ product of 1.05 x 10" T. Hz. From Fig.5 it can be seen that up to this level the variation in core losses is only very small. The minimum value of the parallel loss resistance referred to the 50 Ω level (see Fig.5) is:

$$R_p = \frac{L}{\mu_r} \text{ x } 41 \text{ x } 10^9 = \frac{26.2 \text{ x } 10^{-6}}{100} \text{ x } 41 \text{ x } 10^9 = 10.7 \text{ k}\Omega.$$

So the core losses are: 50/10 700 x 100 % = 0.47 % maximum.

The cable can handle a power of 170 W at 28 MHz and it has a loss of 0.135 dB/m at this frequency. The length of the cable (the sum of L_1 and L_3) is about 60 cm, giving a maximum cable loss of approximately 1.9 %.

When the transformer was driven from an unbalanced 50 Ω source and loaded with two resistors of 6.25 Ω at the balanced side (connecting point of the resistors earthed) the power loss and asymmetry were so small that they could not be measured. The input impedance at the unbalanced side as a function of frequency is given in Table 2 below.

Table 2

f (MHz)	R_p (Ω)	X_p (Ω)
1.6	49.3	+j 395
5	49.6	+j 1250
15	50.2	+j 21200
28	50.2	−j 1060

7.2 50 Ω Unbalanced to 5.55 Ω Balanced Transformer

In this case the transformers described in Sub-Section 6.2 and 6.5 have been combined. The schematic diagram is given in Fig.16. A transformer of this design is again required for a wideband s.s.b. transmitter in the frequency range of 1.6 to 28 MHz. The required peak envelope power handling is 5 W, with an input impedance of 50 Ω unbalanced and an output impedance of 5.55 Ω balanced.

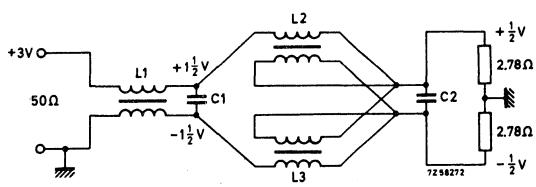


Fig.16. 50 Ω unbalanced to 5.55 Ω balanced transformer.

The transformer can be wound on a single 4C4 toroid of 23 x 14 x 7 mm. Winding L_1 must have a characteristic resistance of 50 Ω; this has been approached by a twisted pair of enamelled copper wires of 0.45 mm diameter (R_c = 52 Ω). Windings L_2 and L_3 must have a characteristic resistance of $16\,^2/_3$ Ω. For this purpose four enamelled copper wires of 0.22 mm diameter have been twisted and diagonally interconnected as shown in Fig.17; the characteristic resistance was found to be 30 Ω, making a high-frequency correction necessary. The required compensation, determined experimentally, gave values for C_1 of 22 pF and for C_2 of 270 pF.

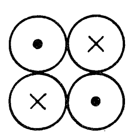


Fig.17. Interconnection of winding wires. In the cross-section shown, the two wires indicated by dots are interconnected, as are those indicated by crosses. This configuration gives a maximum winding capacitance, necessary for a low characteristic resistance.

From the voltage division it follows that winding L_1 must have 1½ times the number of turns of windings L_2 and L_3. The required number of turns can be calculated in the same way as in the previous section. To get a parallel input inductance of 20 μH minimum, 9 turns are required for winding L_1. The measured value was approx. 30 μH.

Calculation of B_{max} according to the method of Sub-Section 7.1 gives a value of approx. 40 gauss at 1.6 MHz. This is so low that a smaller core could be chosen. The reason for not doing this is that the smaller core has not sufficient room for the windings. In Table 3 below, the input impedance is given as a function of frequency.

Table 3

	without h.f. compensation		with h.f. compensation	
	R_p (Ω)	X_p (Ω)	R_p (Ω)	X_p (Ω)
1.6	49	+j 300	48	+j 370
5	50	+j 360	49	+j 840
15	56	+j 220	50	+j 1400
28	71	+j 213	51	−j 11000

If there is no high frequency compensation, the variation in impedance is rather large. This is caused by:

- the characteristic resistance of the transmission lines L_2 and L_3 being 30 Ω instead of $16\,^2/_3$ Ω.
- the electrical length of the windings L_2 and L_3. Measurements have shown that the wavelength on these lines is only 41 % of the wavelength in free space – the decrease in wavelength being caused by the twisting of the wires, the dielectric constant of the wire insulation and the influence of the μ_r of the ferrite core.
- the inductance of the load resistors.

The power loss measured at 28 MHz was 0.34 dB. At lower frequencies it was too small to be measured accurately. Voltage asymmetry between both terminals was less than 4 %.

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ELECTRONIC APPLICATION LABORATORY REPORT

GROUP : C.A.B.
Communication
EINDHOVEN

AUTHOR : M.J. Köppen

Report nr : ECO 7113
Date : 23.7.1971
Project nr: 6009
Pages : S1 + N3 + R3/

TITLE : Single stage wide-band (1.6 - 28 MHz) SSB driver modules with BLY92A and BLX13, operating in class A.

SUMMARY

A description is given of two wideband class-A amplifiers, intended for driver applications in S.S.B. transmitters.
The first amplifier contains a BLY92A and delivers 3W P.E.P. at an I.M.D. of better than -40dB. The gain is 18.1 ±0.2dB over the frequency band of 1.6 to 28 MHz and the input V.S.W.R. is less than 1.3.
The second amplifier uses a BLX13 delivering 8W P.E.P. at an I.M.D. below -40dB. Its gain is 16.8 ±0.2dB over the above mentioned frequency band and the input V.S.W.R. is 1.5 maximum.
Both amplifiers operate from a supply voltage of 28V.

A.H. Hilbers

	Advies Octrooi d.d. 2/9-'71	~~AV~~	GV		B		BL
	Opgave Mamo d.d. 2/9-71	~~AV~~	~~GV~~	~~BI~~	B		BL
	DATUM: 13 aug. 1971	MAMO:					

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Relevant Summary

AMPLIFIER 1

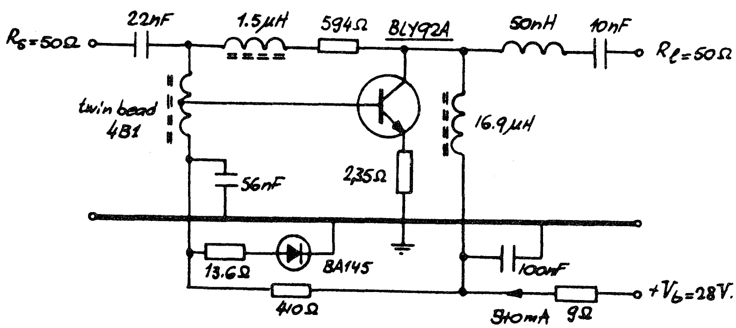


Fig. 1

Advies Octrooi dd: 2/9-'71	~~AV~~	GV		B		BL
Opgave Mamo dd:	AV	GV	EI	B		BL
DATUM: 13 aug. 1971	MAMO:					

AMPLIFIER 2

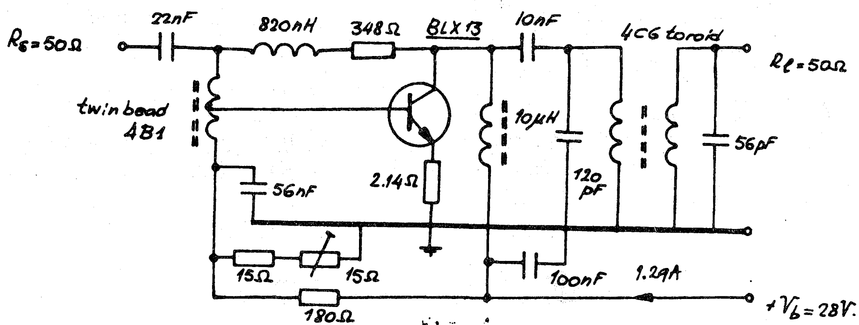


Fig. 4

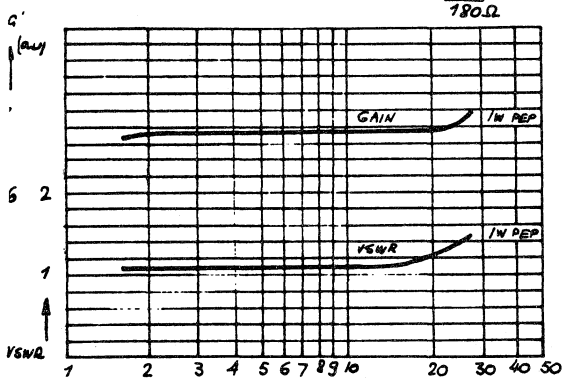


Fig. 5

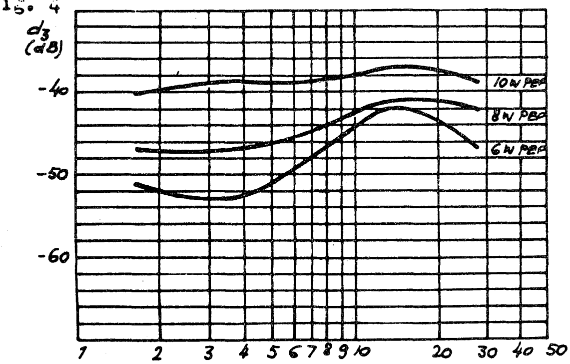


Fig. 6

Fig. 4 shows the circuit of a wide-band SSB module equipped with BLX13 operating in class A. The amplifier covers at least the h.f. band of 1.6 - 28 MHz.

Fig. 5: The behaviour of the power gain and input VSWR versus frequency. Power gain = 16.8 ±0.2dB; VSWR ≤ 1.5.

Fig. 6: The I.M. distortion, characterised by the d_3.

Harmonic suppression:

2nd ≥ 28dB; 3rd ≥35dB at 6 Watts P_o.

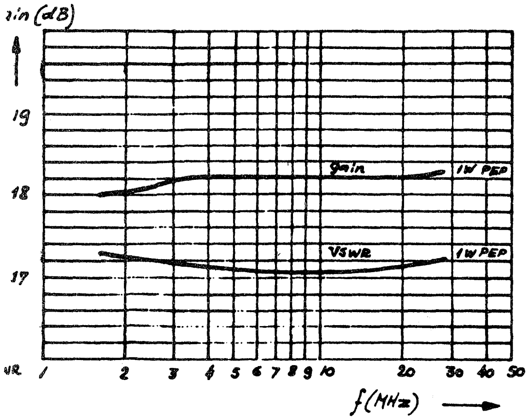


Fig. 2

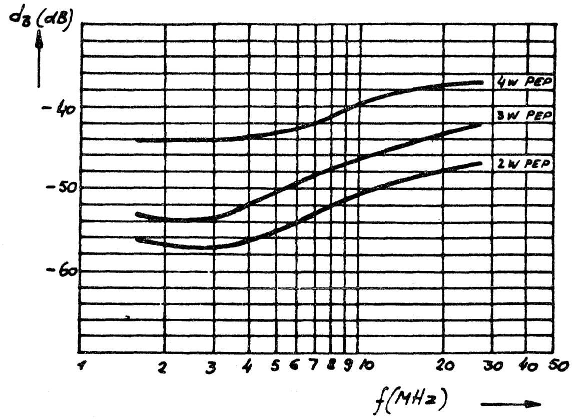


Fig. 3

Fig. 1 shows the circuit of a wide-band SSB module equipped with BLY92A operating in class A. The amplifier covers at least the h.f. band of 1.6 - 28 MHz.

Fig. 2: The behaviour of the power gain and input VSWR versus frequency.
Power gain = 18.1dB $\pm$0.2dB; VSWR $\leq$ 1.3.

Fig. 3: The I.M. distortion, characterised by the d_3.

Harmonic suppression:
2nd $\geq$ 20dB; 3rd $\geq$ 31dB, at 3 Watts P_o.

1. Introduction

C.A.B. report COE 71103 [1] describes a wide-band linear SSB module capable to supply 80-100 watts PEP output covering the main SSB band from 1.6 to 28 MHz.

The amplifier contains two pieces BLX14, connected in push-pull and operating in class AB.
The power gain approximates 16dB, what asks for a drive of abt. 2.5 watts.

This report describes the set-up of two suitable driver modules with resp. BLY92A and BLX13, operating in class A.

The power gain is 17-19dB, resulting in a drive of some tens of milliwatts.

2. Design considerations

The problems occuring with this kind of amplifiers are quite different from those of tuned amplifiers.
Linear and non linear distortion need to be very low over a wide frequency range.

The constancy of the gain versus frequency, the linear distortion, must be within tenths of a dB, whilst the non linear distortion characterized by the intermodulation distance (d_3-d_5) has to be very low. The requirement on I.M. distortion for the complete amplifier is in most countries -26dB or better This can be realized with a final stage having an I.M.D. of -30dB and a driver with -35 to-40dB.
Likewise the input VSWR needs to be small enough over the whole range, because it is directly associated with the loading of a possible pre-driver.
In any case these values need to be superior to those of the final module.

Typical values measured on the final module:

P_o = 100 watts PEP for at least $d_3 \leq -30$dB, whilst the gain is 16dB, what asks for an input power of 2.5 watts.
Highest VSWR is 1.4.

The fact that such a VSWR occurs implies that the driver has to supply a power of at least 2.5 watts into an impedance between 35 and 70 Ohms.

Target:

P_o = 3 watts PEP for $d_3 \leqslant -40$dB corresponding approximately
with: P_o = 4 watts PEP for $d_3 \leqslant -30$dB

Two different drivers have been constructed.

Originally we started with a BLX13 version; afterwards this amplifier appeared too large in capability. However the results will be given in this report as an application example (amplifier 2).

Later an easier set-up has been designed, based on the BLY92A being a 175 MHz transistor (amplifier 1).
It is obvious to start with the latter, having appr. 18dB gain.

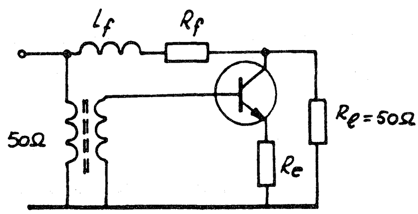
Both amplifiers utilize series- and shunt negative feedback, whilst wide-band transformers serve as matching elements.

In order to obtain the best linearity class A operation has been chosen.

3. Circuit description

3.1 General

In main lines the amplifier is equal to one of those described in ref. [2] [3].
Fig. 1 shows the basic circuit of a feedback amplifier with combined voltage and current feedback.



In- and output impedances are 50Ω.

Fig. 1

To realize the target mentioned in the previous section we must take into account a power loss of appr. 20% in the feedback resistors. So the required transistor output is 5 watts PEP at -30dB I.M.D.
As the maximum efficiency of a class A amplifier is 50%, the required D.C. input power to the transistor must be 10 watts.

3.2 Amplifier 1

The applied transistor BLY92A is a 28 Volts type.

To make an output transformer superfluous the ratio of V_{ce} and I_c must be appr. 50 Ohms. Together with the DC input power of 10 watts this leads to an adjustment of V_{ce} = 22 Volts and I_c ≈450 mA. This operating point lies within the DC SOAR and is acceptable up to heatsink temperatures of 70°C.

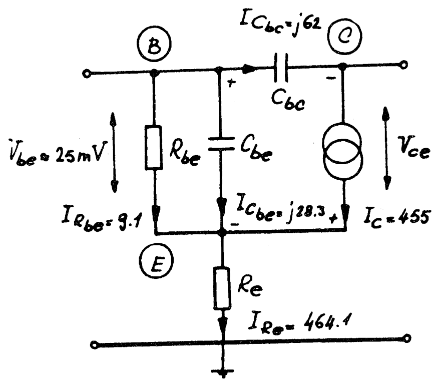


Fig. 2

Fig. 2 shows the simplified HF circuit in which V_{ce} and I_c are in phase.

At V_{cb} = 22 Volts the total collector capacitance is 20.5 pF.
Appr. 71% of this capacitance is between collector and base, so:

$$C_{bc} = 0.71 \times 20.5 = 14.6 \text{ pF.}$$

Due to R.F. excitation the effective value of this capacitance is appr. 10% higher, so it becomes 1.1 x 14.6 = 16 pF.

At the maximum frequency of operation, viz. 28 MHz the reactance will be:

$$X_{bc} = 355 \text{ Ohms.}$$

$$V_{be} = k.T/q \approx 0.025 \text{ Volts at } 25°\text{C.}$$

Because P_{DC} = 10 watts and V_{ce} = 22 Volts, the I_c = 455 mA.
This operating point, related to the f_T curve, is acceptable.

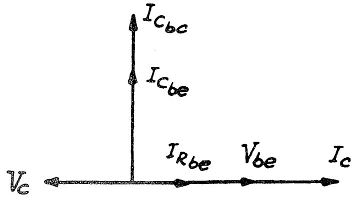
The relation between I_c and $I_{C_{be}}$ is:

$$f_T/f_W = I_c/I_{C_{be}}$$

with f_T typically at 450 MHz (V_{CE} = 22 Volts) and the upper working frequency f_W = 28 MHz $I_{C_{be}}$ follows from:

$$I_{C_{be}} = 0.455 \times 28/450 = 28.3 \text{ mA.}$$

Calculating with h_{FE} = 50, $I_{R_{be}}$ becomes 9.1 mA.
$I_{R_{be}}$ and I_C add in the emitter resistor R_e (Fig. 3).



Hence : I_{R_e} = 455 + 9.1 = 464.1 mA
whilst: $V_{C_{bc}}/-j\,X_{bc}$ = 22.025/-j355 = j 62 mA

Fig. 3

3.3 The input circuit

The transistor can now be inserted in the feedback chain, whilst by means of a transformer the input is adapted to 50 Ohms.
To obtain matching over a wide-bandwidth a 1:4 transmission line type has been applied.

Fig. 4 shows the extended circuit.

The source impedance R = 50 Ohms.

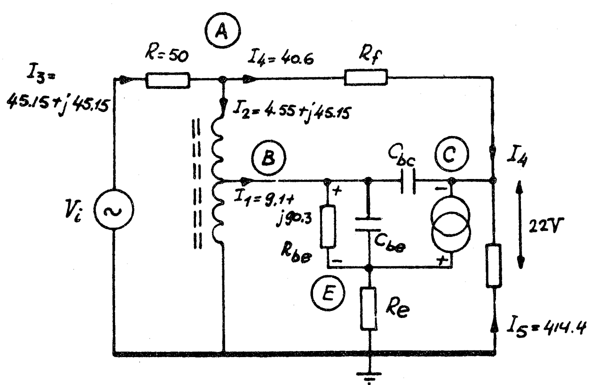


Fig. 4

Because the impedance step is 4:1 the impedance between midtap B and earth is 50/4 = 12.5 Ohms.
The current I_1 has a complex character and is equal to: 9.1 + j 90.3 mA.
So $\overline{I}_2$ = 4.55 + j 45.15 mA.

It is advisable to equalise at point A the real part of the current to the imaginary part so $\overline{I}_3$ = 45.15 + j 45.14 mA. The remainder of the ohmic part 45.15 - 4.55 = 40.6 mA flows into the feedback resistor R_f:
I_4 = 40.6 mA.

Now the input voltage V_A can be derived from

$$R.I_3 = 50 \times 45.15/10^3 = 2.26 \text{ Volts}$$

whilst the effective power becomes:

$$P_{i\ eff} = V_A^2/2.R = 2.26^2/2 \times 50 = 0.051 \text{ Watts.}$$

V_B is equal to half the input voltage: $V_B = 1.13$ Volts causing an emitter voltage of $1.13 - 0.025 = 1.1$ Volt.

R_e can now be derived from:

$$R_e = (V_B - V_{be}) / (I_c + I_b) = 1.1/464.1 = 2.38 \text{ Ohms.}$$

The internal $r_e \approx 0.2$ Ohm , so the external R_e has to be 2.18 Ohms.

3.3.1 The input transformer

Wide-band transformers for the h.f. band can be executed as conventional and transmission line type.

A conventional type consists of a magnetic core and separate primary and secondary windings. In a transmission line type these windings have been combined to form one or more transmission lines, symmetrical or coaxial.

The advantage of the latter is that the stray inductance and the winding capacitance are absorbed in the line configuration so that the upper frequency limit is considerably shifted to higher regions. The primary inductance plays a role at low frequencies.

In this way extremely wide bandwidths can be realized, however, with the disadvantage that only transformation ratios of $(n:m)^2$ e.g. 1:4 are possible. (n and m have to be integers).

In CAB-reports NCO 6814 [2] and ECO 6907 [4] analysis have been given of wide-band transmission line transformers, in the latter with special emphasis on the power handling capability and the compensation technique.

The most suitable core shape was found to be a ferroxcube twin bead as shown in Fig. 5. The 4B1 grade material has been found a good compromise between core losses and obtainable bandwidth.

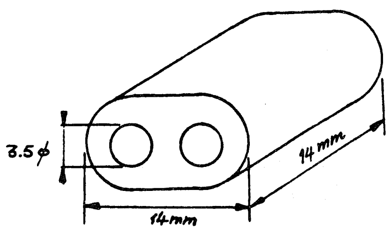
It was found that a line of two twisted CuEm wires of 0.45 mm dia has a characteristic impedance of 51.5 Ohms. This impedance is incorrect when used for a (50:12.5) Ohms transformer - it has to be $(Z_{in} \cdot Z_{out})^{\frac{1}{2}} = 25$ Ohms - as utilised in our case, but in practice this fault is small and can be accepted in this frequency range.

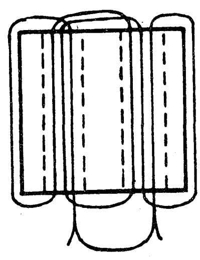
Report NCO 6814 mentions that the power loss of a similar transformer at 27 MHz is:

6% - 5 watts.
4.5% - 14 watts.
3.5% - 40 watts.

whilst the 3rd harmonic distortion, indicating the linearity, is less than 0.1% at 27 MHz for $P_o = 40$ watts.

Some experience has been gained in determining the number of turns and the winding system.
Optimal results were obtained with a type shown in Fig. 6.





2 x 0.45 mm twisted CuEm wire.

Fig. 5

Fig. 6

Catalog number 4312 020 31570

The table below provides some information on the mid-tap impedance when the primary is loaded with 50 Ohms (measurements with the aid of the H.P. 4815A).

f (MHz)	Z (ohms)	Phase (degrees)
1.6	12.80	+9
3.5	13.20	+6.5
7.0	13.25	+9
14.0	13.40	+14.5
28.0	14.60	+25.5

To get an idea of the inductance and coupling factor k the total (L_T) and mid-tap (L_H) values at 1.6 MHz have been measured.

$L_T = 36.2\ \mu H$; $L_H = 9.05\ \mu H$.

Calculating with these values yields an exact value of k = 1. In reality a value of k very close to 1 is attainable. Fig. 7 shows the equivalent circuit at 1.6 MHz.

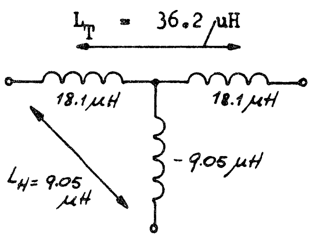


Fig. 7

3.4 The output circuit

With reference to Fig. 4 it can be seen that the collector peak voltage is: $V_{ce} - V_{Re} = 22 - 1.1 = 20.9$ Volts, whilst because $I_c = I_4 + I_5$ the current through R_1 is 455 - 40.6 = 414.4 mA.

According to this R_1 and the power in this resistor can be calculated from:

$$R_1 = V_c/I_5 = 20.9/0.414 = 50.5 \text{ Ohms}$$

$$P_o = V_c \cdot I_5/2 = 20.9 \times 0.414/2 = 4.33 \text{ watts,}$$

what reasonably approaches the target of 4 watts.

3.5 The output matching

In the practical circuit an L-section is inserted between collector and load to compensate for the capacitive output.
The total collector capacitance at V_{cb} = 22 Volts is equal to 20.5 pF.
Due to R.F. excitation this capacitance increases by appr. 10% so it becomes 1.1 x 20.5 = 22.55 pF.

To this value the stud-capacitance of 2 pF must be added, giving 22.55 + 2 = 24.55 pF.
The collector-input feedback impedance contains an inductance (see section 3.6), by which a negative capacitance of 4.3 pF is presented to the collector. Now the total collector-ground capacitance becomes 20.25 pF, having a reactance of -j281 Ohms at 28 MHz.

3.5.1 High frequency compensation

The output resistance of 50.5 Ohms in parallel with the output capacitance have to be adapted to 50 Ohms.
This can be arranged by a simple low-pass section (Fig. 8).

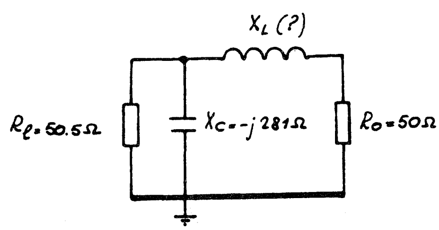


Fig. 8

Very useful wide band matching systems will be found in an article of Horst Nielinger. [5]
With the aid of this article the value of X_L (resp. L) has been determined for the most optimal VSWR.

The calculation process goes as follows:

$$R_o/X_{C\ min} = X_{L\ max}/R_o$$

f_{max} = 28 MHz

$X_{C\ min}$ = $1/2\pi f_{max}.C$ = 281 Ohms

R_o = 50 Ohms

$R_o/X_{c\ min}$ = 50/281 = 0.178 $\longrightarrow$ VSWR max = 1.03

$X_{L\ max}$ = 0.178 x 50 = 8.9 Ohms

L = $X_{L\ max}/2\pi f_{max}$ = 8.9/6.28 x 28 x 10^6 = 50.6 nH.

3.5.2 The collector choke

Some effort has been put in obtaining a suitable collector choke. An acceptable solution appeared to be the application of a 4B1 ferroxcube bead wound with a single layer of 0.25 mm CuEm wire. Between the windings and the bead body a layer of thin transformer paper has been inserted.

The following measurements with the aid of an RF vector impedance meter will show the behaviour over 4 octaves.

f (MHz)	Z (Ohms)	Phase (degrees)
1.6	170	+90
3.5	365	+90
7.0	755	+90
14.0	1850	+90
28.0	43.10^3	+55

Parallel resonance at 28.7 MHz

Inductance at 1.6 MHz is 16.9 μH.

In order to get an idea of the core saturation and its influence on the choke capabilities, some measurements have been made with the set-up of Fig. 9.

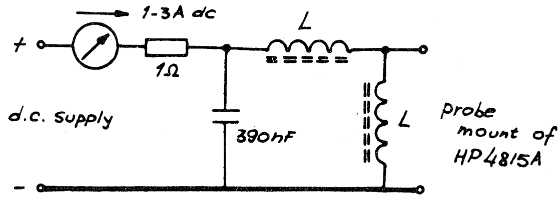


Fig. 9

It appeared that even with 3.5 A the results showed only small deviations.

f (MHz)	Z (Ohms)	Phase (degrees)
1.6	81	+90
3.5	173	+90
7.0	350	+90
14.0	795	+90
28.0	$3{,}25 \cdot 10^3$	+96

The parallel damping R_p at 1.6 MHz has been measured with the Rx meter (HP 250B). The value of one coil is 11.5 kOhms.

3.5.3 Low frequency compensation

For frequencies around 1.6 MHz the influence of the supply choke and the d.c. blocking capacitor is no longer negligible.
Fig. 10 shows how it can be arranged that these components form the elements of a high-pass filter (½ section).

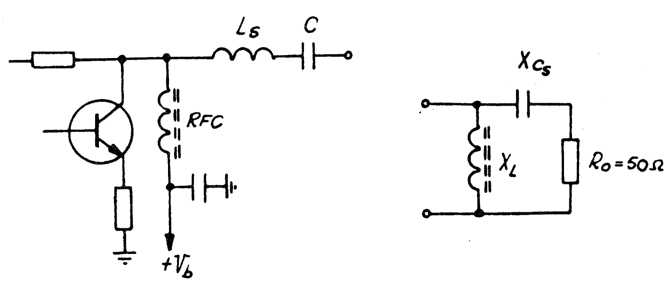


Fig. 10

According to |5 .

$$R_o/X_{L\,min} = X_{C\,max}/R_o$$

f_{min} = 1.6 MHz
$X_{L\ min}$ = $2\pi\ f_{min} \cdot L$ = 170 Ohms
R_o = 50 Ohms
$R_o/X_{L\ min}$ = 50/170 = 0.293 → VSWR max = 1.085
$X_{c\ max}$ = 0.293 x 50 = 14.65 Ohms
$C = 1/2\pi\ f_{min} \cdot X_{c\ max} = 1/6.28 \times 1.6 \times 10^6 \times 14.65 = 6.83$ nF

3.6 Feedback R_f and L_f

The value of the feedback resistor can be derived from the data of Fig. 4.

$$R_f = V_{AC}/I_4 = (20.9 + 2.26)/40.6 \times 10^{-3} = 570 \text{ Ohms.}$$

In practice a parallel connection of 820 and 1800 Ohms has been chosen.

The collector feedback inductance has been determined by experiment. It has been based on favourable experiences with L/R ratios in preceding designs, because the calculation is rather complex.
As Fig. 11 shows one needs to transform the R_{be}, C_{bc}, C_{be} and R_f, in what case shunt compensation has been applied. ($L = \alpha \cdot R^2 \cdot C$).

L_f has been chosen equal to 1.5 µH, being a microchoke.

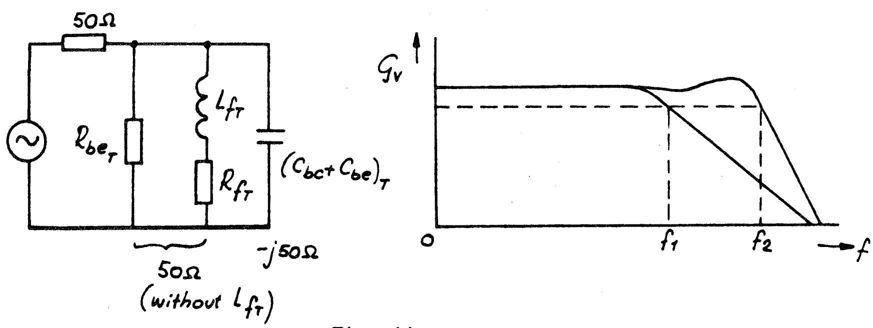


Fig. 11

3.7 D.C. equivalent circuit

As Fig. 12 shows the feedback resistors form part of the d.c. equivalent circuit.

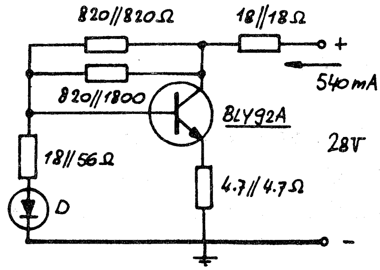


Fig. 12

The component values of Fig. 12 differ somewhat from the calculated ones, however this has almost no consequences for the final results.

To provide a bias voltage that varies with temperature in the same manner as V_{be} of the transistor the circuit contains a compensating diode. To insure a fast thermal response time this diode (BA145) is mounted on the heatsink near the transistor bolt.

The feedback resistors have been composed of several ones in parallel. Particularly the emitter resistor needs to have a small series inductance.

4. Measured results

Measurements have been carried out under the following conditions unless stated otherwise:

V_B = 28 Volts
R_s, R_o = 50 Ohms.
Ambient and stud temperature $T_A = T_S = 25°C$.

4.1 Input VSWR and power gain

These values have been measured versus frequency with the set-up of Fig. 13

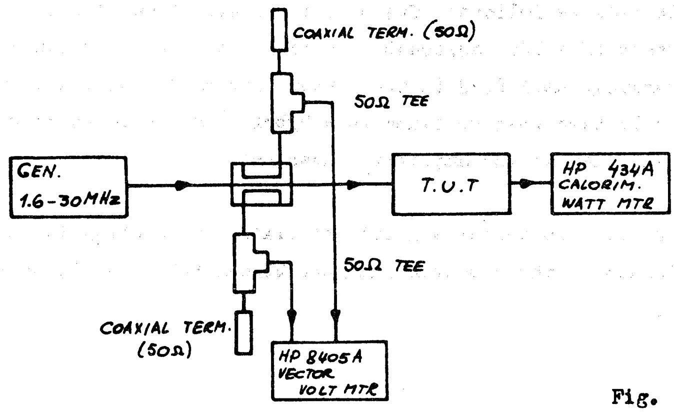


Fig. 13

f (MHz)	$V_{forward}$ (mV)	$V_{refl.}$ (mV)	phase (degr.)	VSWR	gain (dB)
1.6	89	10	72	1.30	18.00
2.0	89	8.8	77	1.22	18.02
3.5	87	4.9	82	1.12	18.20
5	87	3.4	83	1.08	18.20
7	87	2.5	77	1.06	18.20
10	87	2.4	58	1.06	18.20
14	87	3.1	48	1.08	18.20
20	87	5.1	48	1.12	18.20
28	86	8.2	57	1.21	18.30

These measurements have been carried out for P_o = 1 watt.

The results from the vector voltmeter were converted to VSWR and gain with the aid of an HP 9100A calculator. The results are plotted in Fig. 14.

4.2 Intermodulation distortion

The i.m. distortion versus output power is measured with a two tone signal (p, q) at the spot frequencies 1.6, 3.5, 7.0, 14.0 and 28 MHz.

This signal is made as follows: The output signals of two X tal oscillators with a difference of 1 kHz, switchable in pairs at above frequencies, have been separately amplified in two linear wide-band amplifiers. These amplified signals have been combined in a hybrid and via switchable low-pass filters supplied to the amplifier under test.

The oscillator kit also contains X tal oscillators presenting the local oscillator signals to the spectrum analyzer SINGER MF5. Fig. 15 shows the block diagram.

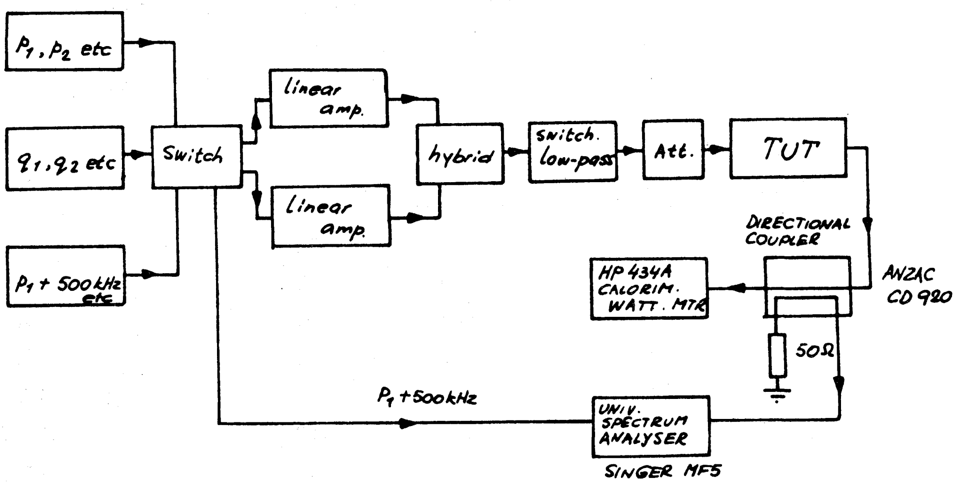


Fig. 15

	P_o = 4 watts PEP		P_o = 3 watts PEP		P_o = 2 watts PEP	
f (MHz)	d_3 (-dB)	d_5 (-dB)	d_3 (-dB)	d_5 (-dB)	d_3 (-dB)	d_5 (-dB)
1.6	44	55	53	>60	56	>60
3.5	44	53	53	>60	57	>60
7	42	53	48	>60	53	>60
14	38	55	45	>60	49	>60
28	37	55	42	>60	47	>60

4.3 Harmonic suppression

Fig. 17 shows the spectrum of the in-band harmonic content for a two-tone signal p, q.

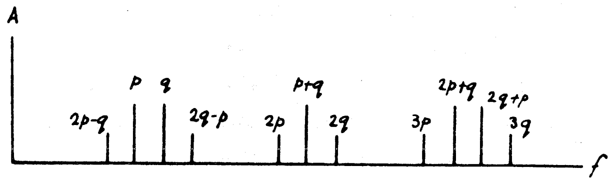


Fig. 17

Measurements have been done with single tone signals (p) of 1.6, 3.5, 7, 14 and 28 MHz, being already available in the two-tone equipment.

The block diagram of the measuring set-up is depicted in Fig. 18.

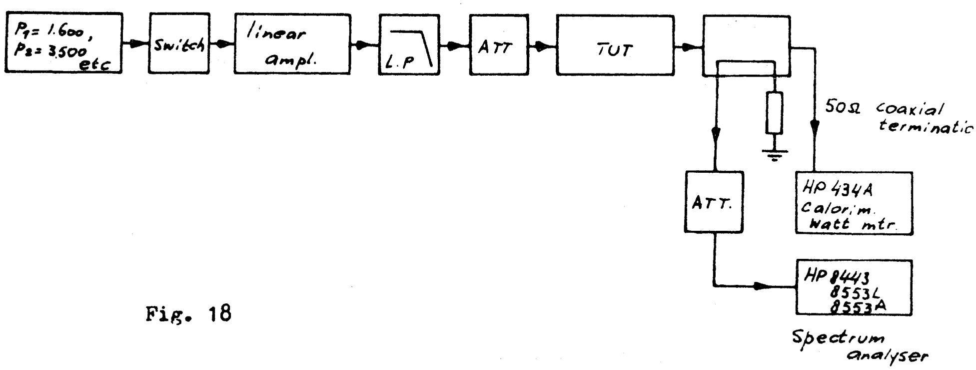


Fig. 18

P_o = 3 watts

f :	1.6 MHz	3.5 MHz	7 MHz	14 MHz	28 MHz
2^{nd} harmonic (2p):	-44dB	-39dB	-33dB	-27dB	-20dB
3^{rd} harmonic (3p):	-48dB	-43dB	-37dB	-31dB	-33dB

5. <u>Amplifier 2</u>

The calculation of this amplifier is in main lines similar to the foregoing. So it shall be given briefly.
Because the output impedance of the applied BLX13 has a lower value a wide-band toroid transformer has been applied.
The target for this amplifier was to reach maximum possible output power at a supply voltage of 28 Volts and a heatsink temperature of 70°C maximum. As the voltage drop across the emitter resistor is appr. 2 Volts we chose V_{ce} = 26 Volts. At this voltage and T_{hs} = 70°C a D.C. dissipation of 28 watts is allowed in the transistor.
So I_c = 28/26 = 1.075 A and the maximum transistor output power is 50% of 28 watts or 14 watts. The maximum output power of the circuit is appr. 20% less due to the loss in the feedback resistors, so it becomes 11 to 12 watts at an I.M.D. of -30dB, which corresponds with 8 to 9 watts at -40dB I.M.D.

C_{be} = 28.4 pF; this value has been obtained in the same way as in section 3.2. Fig. 19 shows the simplified equivalent circuit with V_{ce} and I_c in phase.

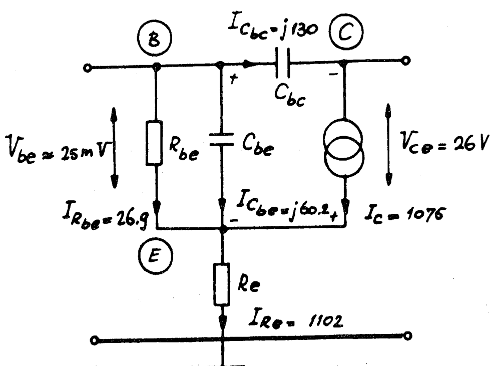


Fig. 19

V_{be} = k.T/q ≈ 25 mV.

f_T = 500 MHz typ.

f_W = 28 MHz

I_{Cbe} = $I_c \cdot f_W/f_T$ = 1.075 x 28/500 = 60.2 mA

h_{FE} = 40; $I_b = I_c/h_{FE}$ = 1.075/40 = 26.9 mA

I_{R_e} = $I_c + I_b$ = 1075 + 26.9 = 1102 mA

$-jX_{cb}$ at 28 MHz = -j 200 Ohms.

I_{Cbc} = $V_{Cbc}/-j\,X_{cb}$ = 26.025/-j200 = j 130 mA

Fig. 20 shows the extended circuit with feedback elements.

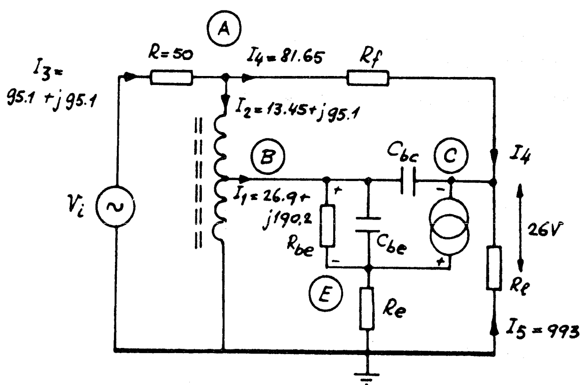


Fig. 20

The input transformer is a 4:1 transmission line type. For details see amplifier 1.

The source resistance is 50 Ohms, so the midtap impedance is 50/4 = 12.5 Ohms.

$\bar{I}_1$ = 26.9 + j 190.2 mA and $\bar{I}_2$ = 13.45 + j 95.1 mA.

$\bar{I}_3$ = 95.1 + j 95.1 mA

The remainder of the real part flows into the R_f chain: I_4 = 81.65 mA.

V_A = $R.I_{3R}$ = $50 \times 95.1/10^3$ = 4.755 Volts

$P_{i\ eff}$ = $V_A^2/2.R$ = $4.755^2/2.50$ = 0.226 Watt.

$V_B = \frac{1}{2}V_A = 4.755/2 = 2.377$ Volts.

$V_{R_e} = V_B - V_{be} = 2.377 - 0.025 = 2.352$ Volts
$R_e = V_{Re}/I_{Re} = 2.352/1.102 = 2.13$ Ohms
Internal $r_e = 0.22$ Ohm so external R_e has to be 1.91 Ohm.

The collector peak voltage is $V_{ce} - V_{Re} = 26 - 2.352 = 23.65$ Volts.
$I_5 = I_c - I_4 = 1.075 - 81.65 = 993$ mA.

$R_1 = V_c/I_5 = 23.65/0.993 = 23.75$ Ohms
$P_o = V_c.I_5/2 = 23.65 \times 0.993/2 = 11.75$ Watts.

The total output capacitance, calculated in the same way as in section 3.5, is 45.4 pF corresponding with a reactance of -j125 Ohms at 28 MHz.

Because the output circuit consists of the parallel connection of 23.75 Ohms and 45.4 pF it is advisable to match via a compensated transformer.

5.1 The output transformer

In order to prevent difficulties with the set-up of a wide-band transmission line type transformer with a ratio $T = (n_{sec}/n_{prim})^2 \approx 50/23.75 \approx 2$ a conventional type with separate windings has been chosen.

In practical realisation the transformer consists of a 4C6 ferroxcube toroid wound with two separate windings $n_{prim} = 12$ turns and $n_{sec} = 17$ turns i.e. $(17/12)^2 = 2$.

To prevent an unacceptable spreading inductance each winding has been distributed over the whole core length, consequently the primary has been wound in between the secondary (Fig. 21).

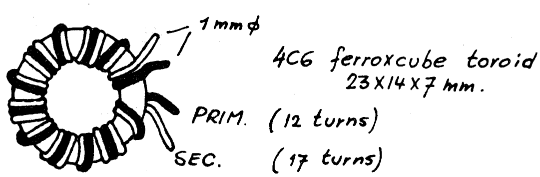
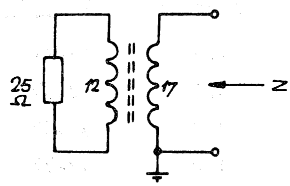


Fig. 21

With the aid of a vector impedance meter (HP-4815A) it can be seen that the behaviour deviates in the upper frequency range. Measurements were made with an ohmic primary termination of 25 Ohms.

f (MHz)	Z (Ohms)	Phase (degrees)
1.6	50.5	12
3.5	51.0	11
7.0	52.5	14
14.0	56.5	24
28.0	70.5	40



Spreading inductance at 1.6 MHz measured secondary is 232 nH.

5.2 High frequency compensation

Fig. 22 shows the compensation system seen from the secondary. R_o is the transformed output resistance of the transistor and L the transformed spreading inductance seen from the secondary. According to Nielinger [5] two capacitors $C_1 = C_2$ will be added to form a low-pass section by which the max. VSWR is reduced.
C_1 is transformed from the primary and consists partly of the transistor output capacitance.

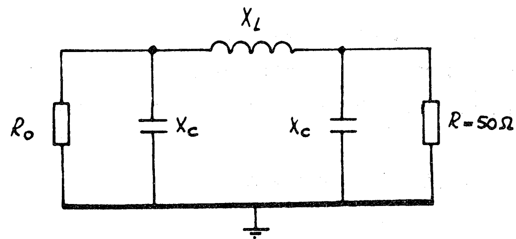


Fig. 22

$X_{L\,max} = 2\pi\ f_{max} \cdot L = 42.5$ Ohms
$f_{max} = 28$ MHz
$X_{L\,max}/R = 42.5/50 = 0.85 \longrightarrow$ VSWR max $= 1.06$
From the curve in [5] it follows that:
$R/X_{c\,min} = 0.515$ $\qquad X_{c\,min} = 50/0.515 = 97$ Ohms
$X_{c\,min} = 1/2\pi\ f_{max} \cdot C$
$C = 1/6.28 \times 28 \times 10^6 \times 97 = 58.5$ pF
Really the latter is secondary capacitance, whilst the primary becomes
T. C $= 2 \times 58.5 = 117$ pF.

The compensated transformer (Fig. 22) has been measured with the vector impedance meter. First the parasitic properties of the mount have been determined. The results have been put in a small digital computer, which output shows the performance versus frequency given in R_s; X_s and VSWR.

f (MHz)	R_s (Ohms)	X_s (Ohms)	VSWR
1.6	49.91	7.78	1.17
3.5	50.39	2.22	1.05
7.0	50.42	0.01	1.01
14.0	49.92	-0.88	1.02
28.0	50.86	-1.65	1.04

The compensated unit couples the transistor with the 50 Ohms load impedance. Although the primary capacitor of 120 pF has to be decreased with the already present capacitance it appeared to be superfluous to do this with respect to linearity and gain; even the I.M.D. at lower levels became slightly better.

5.2.1 The collector choke

Constructively the choke is equal to the one described in 3.5.2.
Only the number of turns differs. (24 turns; 0.35 mm CuEm wire).
The inductance at 1.6 MHz is 10.4 µH (j105 Ohms), whilst an I_{DC} up to 3 A showed no problems.
The parallel damping at 1.6 MHz R_p = 2.3 kOhms.

5.3. Low-frequency compensation

Fig. 23 shows the essential elements for a high-pass section at 1.6 MHz.

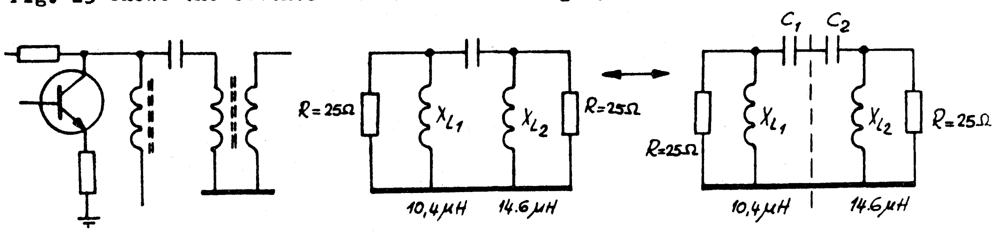


Fig. 23

The primary inductance of the transformer at $f = 1.6$ MHz ($\bar{Z} = 25.7$ Ohms; $\varphi = 10°$) is equal to $14.6\,\mu H$ (j147 Ohms).
Because the filter section is asymmetrical, it has been split up, so:
$C = C_1 . C_2/(C_1 + C_2)$.
According to [5]

$$R/X_{L1\ min} = X_{C1\ max}/R$$
$$X_{L1\ min} = 105 \text{ Ohms}$$
$$R/X_{L1\ min} = 25/105 = 0.238 \longrightarrow \text{VSWR} = 1.07$$
$$X_{C1\ max} = 0.238 \times 25 = 5.95.$$
$$C_1 = 1/\ 2\pi f_{min} .\ X_{C1\ max} = 1/6.28 \times 1.6 \times 10^6 \times 5.95 = 16.7 \text{ nF}$$

and:

$$R/X_{L2\ min} = X_{C2\ max}/R$$
$$X_{L2} = 147 \text{ Ohms}$$
$$R/X_{L2\ min} = 25/147 = 0.17 \longrightarrow \text{VSWR} = 1.025$$
$$X_{C2\ max} = 0.17 \times 25 = 4.25 \text{ Ohms}$$
$$C_2 = 1/2\pi\ f_{min} .\ X_{c2\ max} = 1/6.28 \times 1.6 \times 10^6 \times 4.25 = 23.3 \text{ nF}$$

$C = (16.7 \times 23.3)/(16.7 + 23.3) = 9.75$ nF   10 nF has been chosen.

5.4 Feedback R_f and L_f

$R_f = V_{AC}/I_4 = (23.65 + 4.755)/\ 81.65 \times 10^{-3} = 348$ Ohms.
In practice a parallel connection of 3 times 1.2 kOhms and one 2.7 kOhms resistors has been chosen.

According to reasons argued under 3.6 the value of $R_f = 1.5\,\mu H$.
A microchoke of this value has been applied.

5.5 D.C. equivalent circuit

Fig. 24 shows the d.c. equivalent circuit.

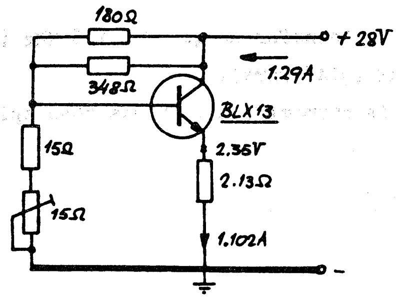


Fig. 24

The 15 Ohms pot. meter has to be adjusted to I_T = 1.29 A for a supply voltage of 28 Volts. As the parts list shows the feedback resistors have been composed of several ones connected in parallel.

6. Measured results (amplifier 2)

Conditions: see 4.

6.1 Input VSWR and power gain

Set-up: Fig. 13

f (MHz)	$V_{forward}$ (mV)	$V_{refl.}$ (mV)	phase (degr.)	VSWR	gain (dB)
1.6	104	4.3	98	1.086	16.65
2.0	103	4.1	88	1.083	16.73
3.5	103	3.7	67	1.075	16.73
5	103	3.6	60	1.072	16.73
7	103	3.5	58	1.070	16.73
10	103	3.7	62	1.075	16.73
14	103	4.8	75	1.098	16.73
20	103	9.5	80	1.203	16.73
28	100	20.3	62	1.509	16.99

These measurements have been carried out for P_o = 1 watt.
The results are plotted in Fig. 25.

6.2 Intermodulation distortion

Set-up: Fig. 15.

	P_o = 10 watts PEP		P_o = 8 watts PEP		P_o = 6 watts PEP	
f (MHz)	d_3 (-dB)	d_5 (-dB)	d_3 (-dB)	d_5 (-dB)	d_3 (-dB)	d_5 (-dB)
1.6	40	50	47	> 60	51	>60
3.5	39	47	47	58	53	>60
7	39	47	45	58	48	>60
14	37	51	41	62	42	>60
28	39	53	42	58	47	>60

The results are plotted in Fig. 26.

6.3 Harmonic suppression

Set-up: see Fig. 18

P_o = 6 watts

f :	1.6 MHz	3.5 MHz	7 MHz	14 MHz	28 MHz
2^{nd} harmonic (2p)	-38dB	-38dB	-31dB	-28dB	-29dB
3^{rd} harmonic (3p)	-46dB	-43dB	-40dB	-35dB	-59dB

7. Mechanical lay-out

Figs. 27 and 28 respectively show the drawings of the p.c.-board for amplifier 1 and 2 with the situation of the components.

The lower sheet of the double clad p.c. board functions as a ground plane. Interconnections of some upper parts with the ground were made with 2 mm tubular rivets. These rivets were soldered to the print conductors to be sure of contact.

Because the amplifiers contain emitter resistors a part of the ground plane has been insulated.

During measurements the p.c. board is placed on a water-cooled heatsink.

8. References

1. J. Mulder : "Short Notes on an 80-100 W PEP Linear Power Amplifier with two pieces BLX14 in the Frequency Band of 1.6-28 MHz". CAB report COE 71103.

2. J.M. Siemensma: "Investigations on linear power amplifiers for SSB. Wideband class A operation from 1.6 to 28 MHz". CAB report NCO 6814.

3. J. Mulder : "A Linear Solid-State Wideband Driver for Medium Power S.S.B. Tubes (f = 1.6 - 28 MHz)". CAB report ECO 6917.

4. A.H. Hilbers : "On the design of H.F. wideband power transformers". CAB report ECO 6907.

   or : "Design of H.F. wideband Power transformers". Application Information 530.

5. H. Nielinger : "Optimale Dimensionierung von Breitband-anpassungs-netzwerken". NTZ Heft 2 1968 pp. 88-91.

M.J. Köppen

/AB 23.7.1971

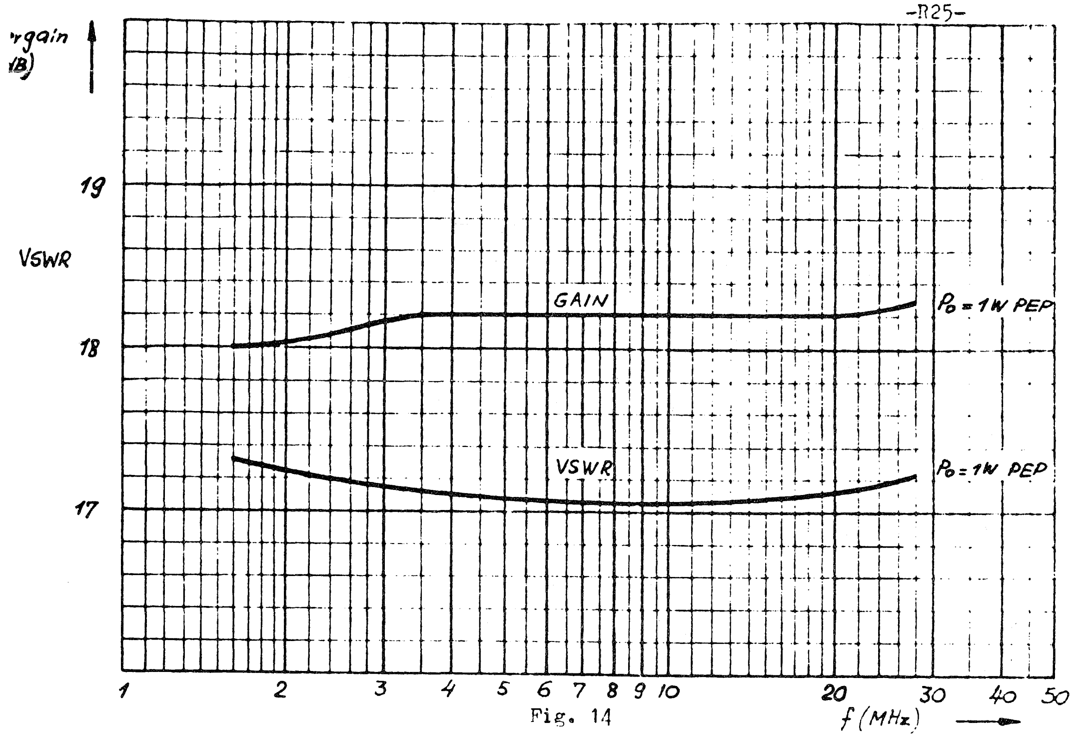

gain
(dB)
19
VSWR
18
17
GAIN
VSWR
P_o = 1W PEP
P_o = 1W PEP
1
2
3
4
5
6
7
8
9
10
20
30
40
50
f (MHz)

Fig. 14

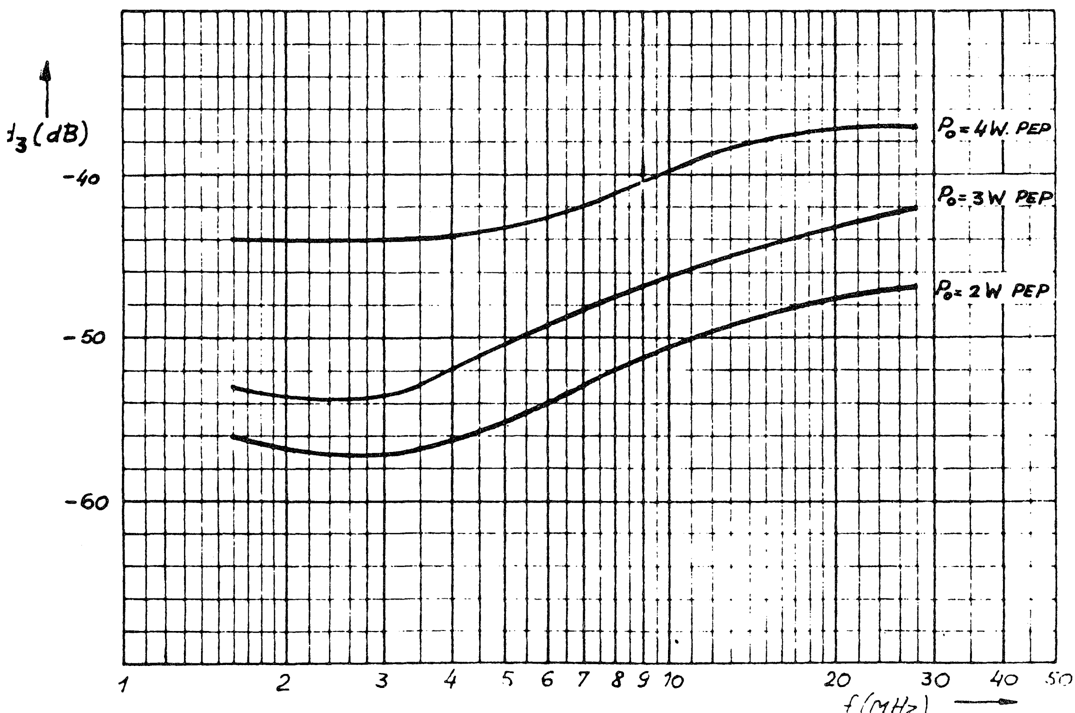

d_3 (dB)
-40
-50
-60
P_o = 4W PEP
P_o = 3W PEP
P_o = 2W PEP
1
2
3
4
5
6
7
8
9
10
20
30
40
50
f (MHz)

Fig. 16

AMPLIFIER 1

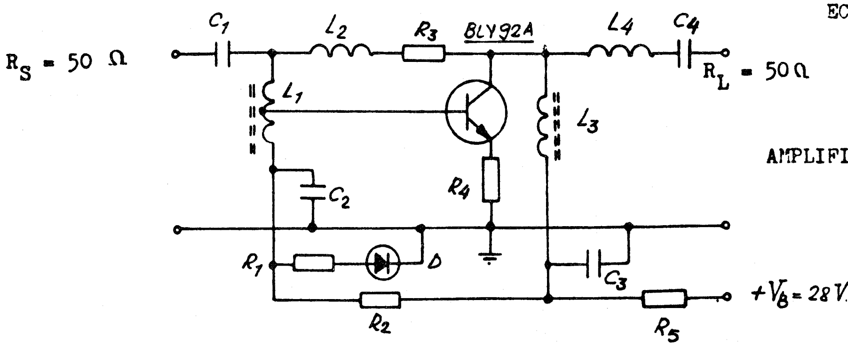


Fig. 29

Components:

R_1	=	parallel connection of:		
		1 x 18 Ohms, carbon ±5%	CR25 style	2322 101 33189
	and	1 x 56 Ohms, carbon ±5%	CR25 style	2322 101 33569
R_2	=	parallel connection of:		
		2 x 820 Ohms, carbon ±5%	CR52 style	2322 101 63821
R_3	=	parallel connection of:		
		1 x 820 Ohms, carbon ±5%	CR52 style	2322 101 63821
		1 x 1.8 kOhms, carbon ±5%	CR52 style	2322 101 63182
R_4	=	parallel connection of:		
		2 x 4.7 Ohms, carbon ±5%	CR25 style	2322 101 33478
R_5	=	parallel connection of:		
		2 x 18 Ohms, carbon ±5%	CR68 style	2322 214 13189
C_1	=	22 nF polyester ±10%		2222 342 45223
C_2	=	56 nF polyester ±10%		2222 342 45563
C_3	=	100 nF polyester ±10%		2222 342 45104
C_4	=	10 nF polyester ±10%		2222 342 45103

L_1 = 14 mm ferroxcube 4B1 twinbead, 4312 020 31520
with 3 turns of 2 x 0.45 mm twisted CuEm wire (see Fig. 6).

L_2 = microchoke 1.5 μH 2422 535 00158

L_3 = ferroxcube bead 4B1, 4312 020 31550
with 33 turns of 0.25 mm CuEm wire, close wound.
L = 16.9 μH, R(DC) < 0.48 Ohms, R_p > 11.5 kOhms.

L_4 = 45 nH, 6 turns closely wound, D_{int} = 6 mm, d = 1 mm CuEm wire, leads 2 x 5 mm.

D = BA145.

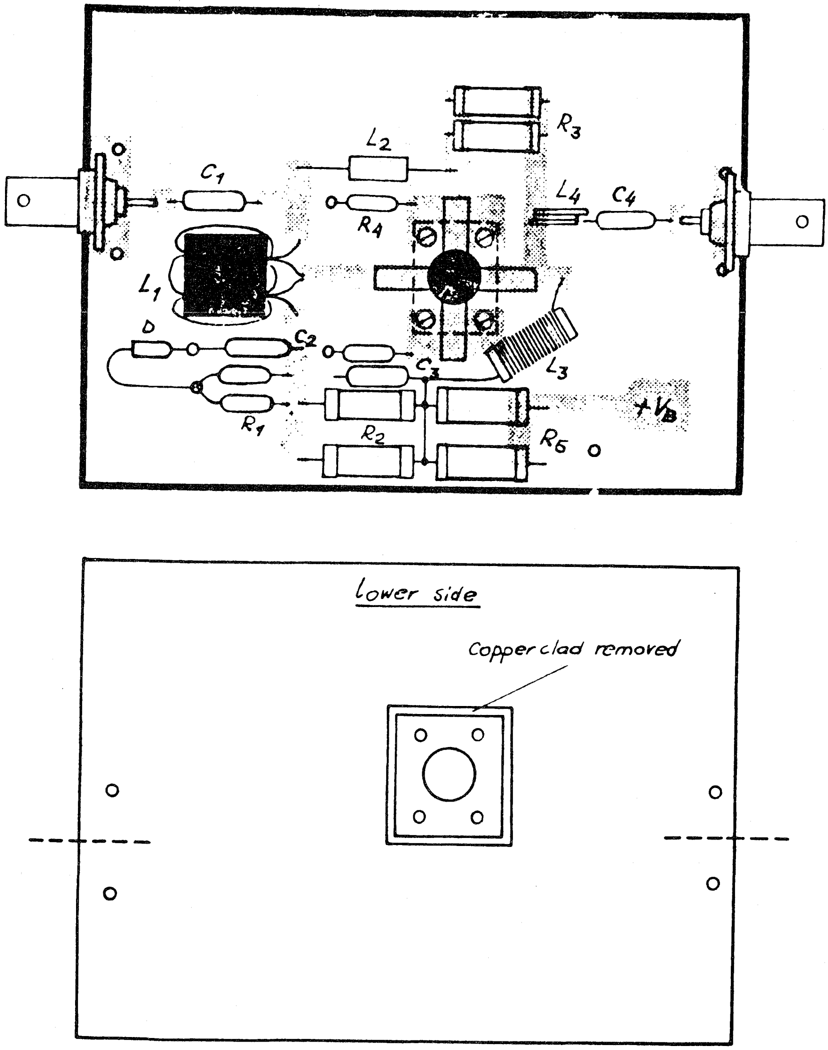


Fig. 27 scale 1:1

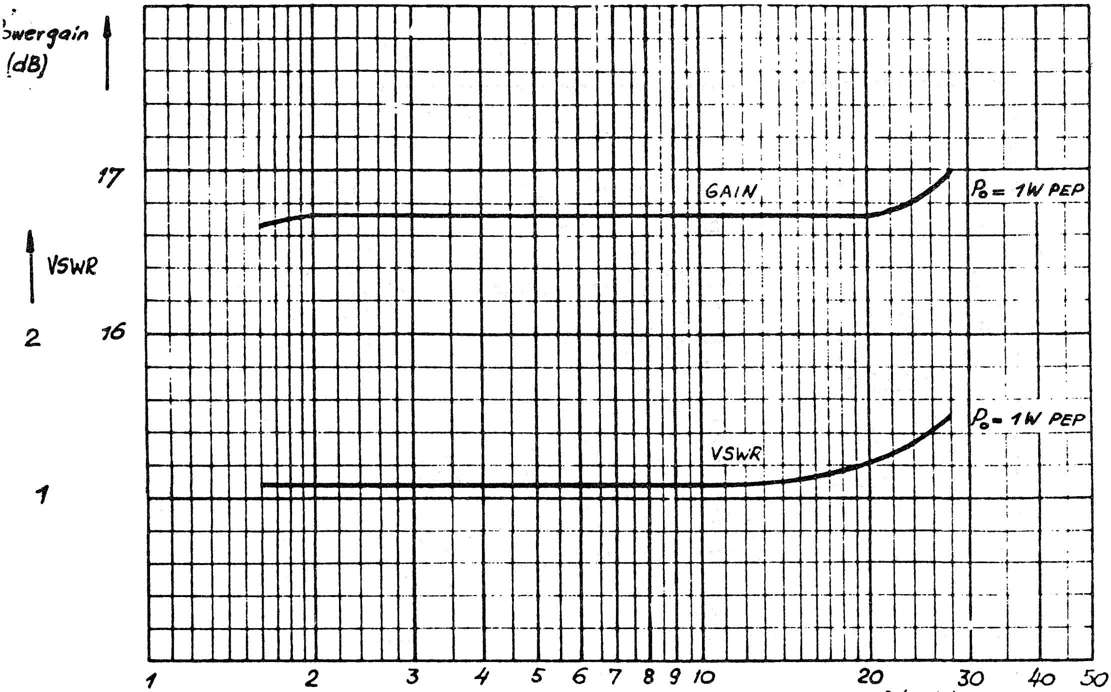

owergain (dB)
VSWR
17
16
2
1
GAIN
P_o = 1W PEP
VSWR
P_o = 1W PEP
1
2
3
4
5
6
7
8
9
10
20
30
40
50
f (MHz)

Fig. 25

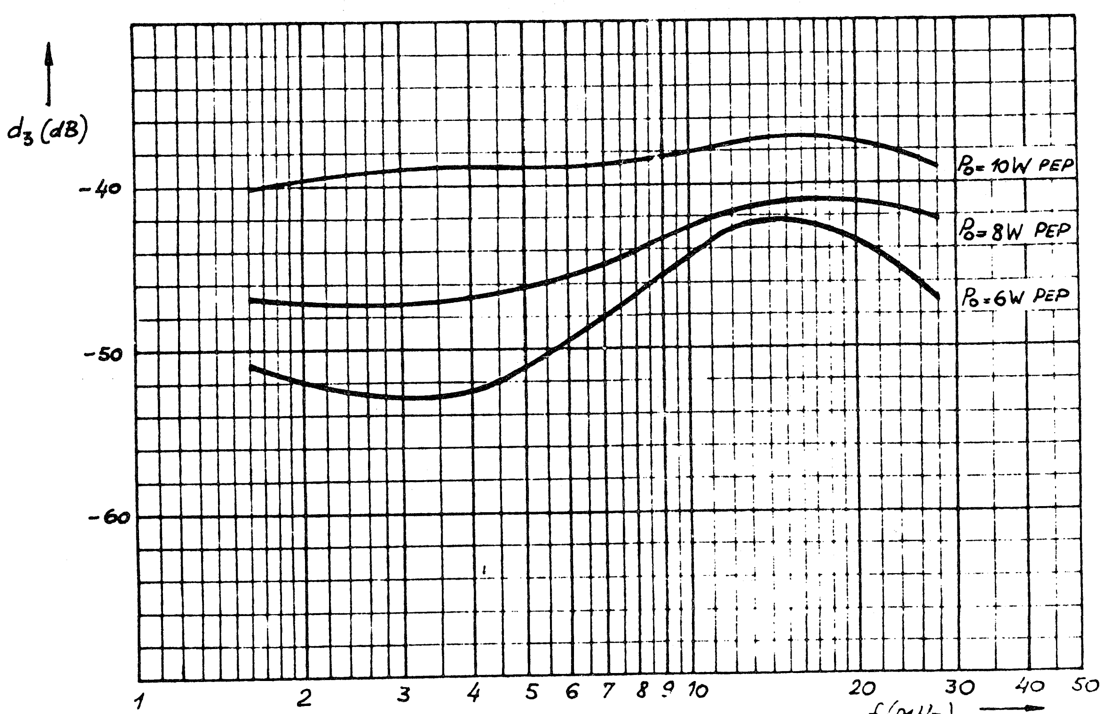
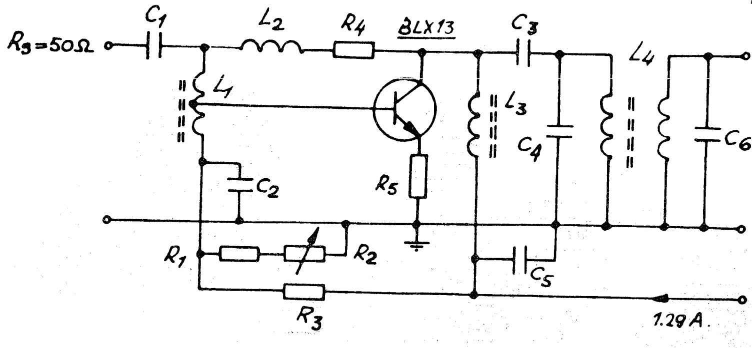

d_3 (dB)
-40
-50
-60
P_o= 10W PEP
P_o= 8W PEP
P_o= 6W PEP
1
2
3
4
5
6
7
8
9
10
20
30
40
50
f (MHz)

Fig. 26

ECO 7113
-R29-



AMPLIFIER 2

Fig. 30

Components:

R_1	=	15 Ohms carbon $\pm 5\%$ CR68 style	2322 214 13159
R_2	=	15 Ohms wire-wound trimming pot.meter, 2 Watts	2322 011 02159
R_3	=	180 Ohms enamelled wire-wound, $\pm 10\%$, 5.5 Watts	2322 320 31181
R_4	=	parallel connection of:	
		3x 1.2 kOhms carbon $\pm 5\%$ CR37 style	2322 212 13122
and		1x 2.7 kOhms carbon $\pm 5\%$ CR37 style	2322 212 13272
R_5	=	parallel connection of:	
		7 x 15 Ohms carbon $\pm 5\%$ CR37 style	2322 212 13159
C_1	=	22 nF polyester $\pm 10\%$	2222 342 45223
C_2	=	56 nF polyester $\pm 10\%$	2222 342 45563
C_3	=	10 nF polyester $\pm 10\%$	2222 342 45103
C_4	=	120 pF ceramic	2222 555 56121
C_5	=	100 nF polyester $\pm 10\%$	2222 342 45104
C_6	=	56 pF ceramic	2222 555 56569
L_1	=	14 mm ferroxcube 4B1 twinbead, with 3 turns of 2 x 0.45 mm twisted CuEn wire (see Fig. 6)	4312 020 31520
L_2	=	820 nH, 15 turns, D_{int} = 6 mm, d = 0.7 mm CuEn wire, close wound.	
L_3	=	ferroxcube bead 4B1 with 25 turns of 0.35 mm CuEn wire, close wound. L = 10 μH, R(DC) < 0.24 Ohms, R_p > 2.5 kOhms.	4312 020 31550
L_4	=	4C6 ferroxcube toroid 23 x 14 x 7 mm; prim.: 12 turns 1 mm CuEn wire; sec. : 17 turns 1 mm CuEn wire (24 μH typ.) (see Fig. 20).	4322 020 91070

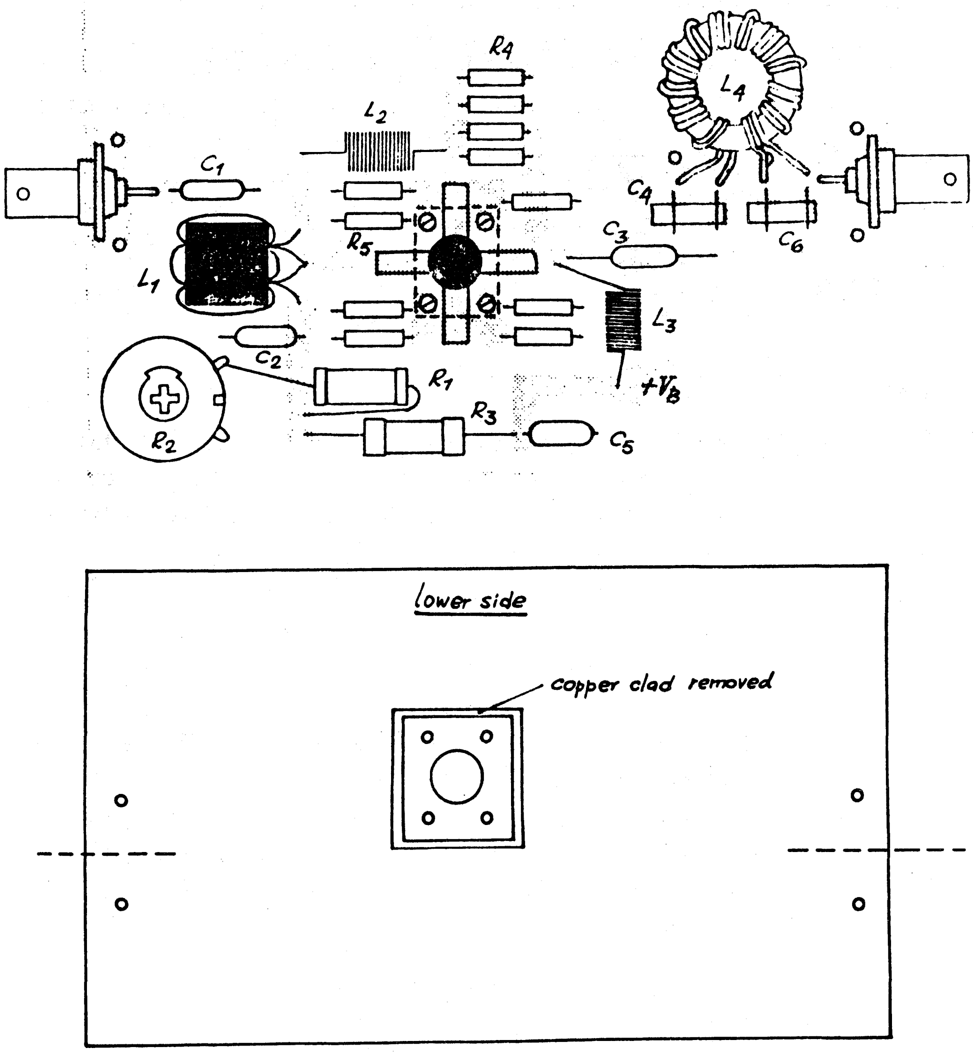


Fig. 28 scale 1:1

ELECTRONIC APPLICATION LABORATORY REPORT

GROUP : C.A.B.
Communication
EINDHOVEN

Report nr : ECO 7201
Date : 27.1.1972
Project nr : 6011
Pages : S1 + M2 + R31

AUTHOR : M.J. Köppen

TITLE : Wide-band (1.6 - 28 MHz) driver stage for the SSB tube YL1230, equipped with two BLX13's in class A.

SUMMARY

This driver stage, which is equipped with 2 pieces BLX13 in push-pull, is intended to drive the 1 kW S.S.B. tube type YL1230. Both transistors operate in class-A from a supply voltage of 28 V. The circuit contains cross-neutralisation and is wideband (1.6 - 28 MHz).
The required drive power for the transistors is 142 mW P.E.P. ± 0.7dB and the input V.S.W.R. is below 1.6. The input voltage for the tube has an I.M.D. ≤ -45dB.
The driver can also be used separately (on a 50 ohm basis) and then it is able to deliver 14 W P.E.P. at an I.M.D ≤ -41dB. In that case the gain is 18.7 ± 0.15dB and the input V.S.W.R. is below 1.5.

A.H. Hilbers

Advies Octrooi d.d: 14/2-'72	~~AV~~	GV		B		BL
Opgave Mamo d.d 17-2-'72	~~AV~~	~~GV~~	~~K~~	B		BL
DATUM: 1 feb. 1972	MAMO:					

ELECTRONIC APPLICATION LABORATORY REPORT

GROUP : C.A.B.
Communication
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AUTHOR : M.J. Köppen

TITLE : Wide-band (1.6 - 28 MHz) driver stage, for the SSB tube YL1230, equipped with two BLX13's in class A.

Relevant Summary

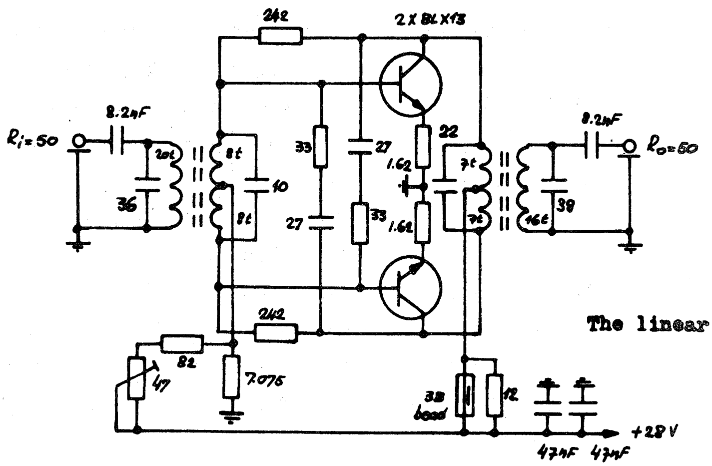


Fig. 1

The linear amplifier.

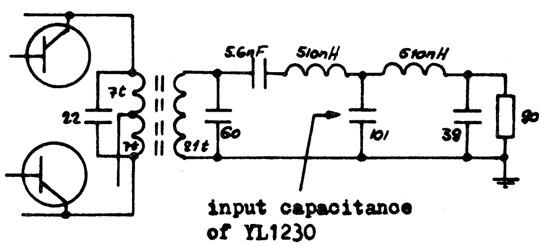


Fig. 2

Extension to tube driver.

	Advies Octrooi d.d: 14/2-'72	~~AV~~	GV		B		BL
	Opgave Mamo d.d. 17/2-'72	~~AV~~	~~GV~~	~~BI~~	B		BL
	DATUM: 1 feb 1972	MAMO:					

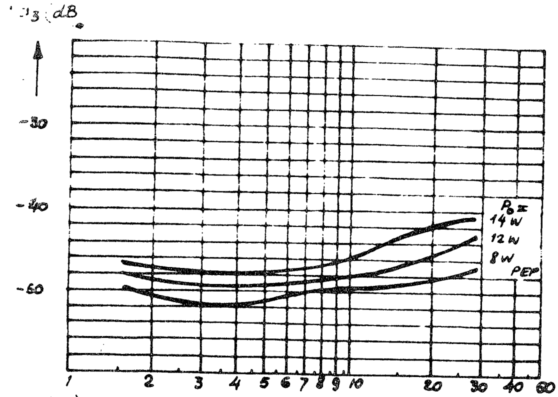


Fig. 3.

The I.M.D., characterised by the d_3, of th
linear amplifier of Fig. 1.
Except for the point f = 28 MHz, P_o = 14 W
PEP where d_5 = -58dB, the d_5 is better tha
-60dB.

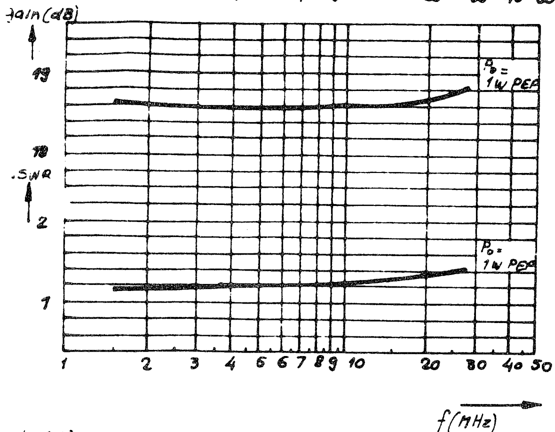


Fig. 4.

The behaviour of the power gain and input VSWR versus frequency.
Power gain = 18.69 ± 0.13dB
VSWR ⩽ 1.43
Harmonics suppression:
2nd ⩾ 28dB, 3rd ⩾ 14dB at 14 Watts P_o

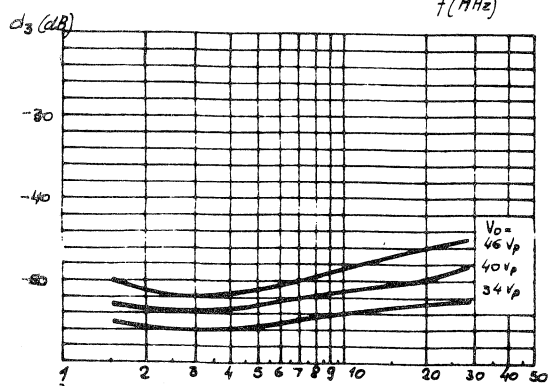


Fig. 5.

The I.M.D. characterised by the d_3, of the tube driver (Figs. 1 and 2).
The d_5 is better than -60dB.

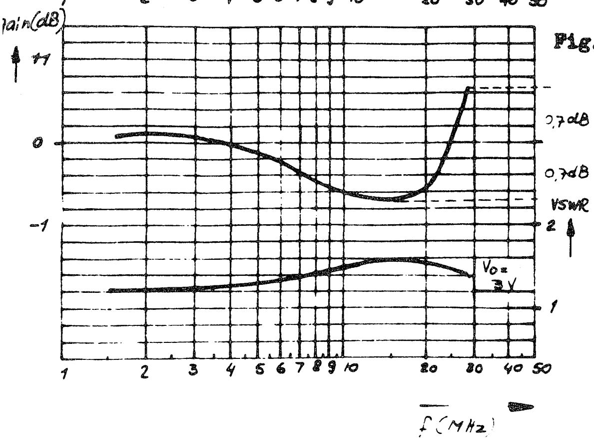


Fig. 6.

The behaviour of the relative voltage gain versus frequency.
Gain variation ± 0.70dB.
VSWR ⩽ 1.58
Harmonics suppression:
2nd ⩾ 42dB, 3rd ⩾ 37dB for 46 Volts pk.

1. Introduction

This report contains the description and measured results of a single stage push-pull broad-band power amplifier, operating in class A.

The requirement was to construct a linear amplifier able to drive the tube YL1230 up to 1 kW PEP output over the h.f. band 1.6 - 28 MHz.

Under full operating conditions the total I.M.D. products of 3^{rd}, 5^{th} and 7^{th} order have to be at least 36dB down, what implies an IMD of appr. 40dB down for the mentioned amplifier.

To adapt the tube input capacitance (grid driven) a simple low-pass filter is needed.

As an alternative, it is possible to apply the amplifier without filter, on 50 Ohms output, as a driver for a 300 Watts PEP amplifier.

The power gain is appr. 19dB.

2. Design considerations

It is obvious to start with the 50 Ohms design, after which the filter unit shall be added to complete the tube driver.

The recently published CAB-report ECO 7114 [1] argues the advantages of applying the push-pull configuration with cross neutrodynisation. This method has been chosen.

The tube driver has to supply 46 Volts peak over 101 pF. It can be calculated that this corresponds with P_o = 10-12 Watts PEP to the characteristic impedance of the filter.

Based on the results with an amplifier described in CAB-report ECO 7113 [2] a pair of BLX13 in class A can do the job. (BLX13 in class A: P_o = 8 Watts PE IMD better than 40dB.)

As remarked the tube represents a capacitance of 101 pF. A favourable adaption over a wide-frequency range can be realised with an L-P filter. The component values have been obtained with the aid of a computer program.

3. Circuit description

3.1 Basic circuit

Fig. 1 shows the basic circuit.

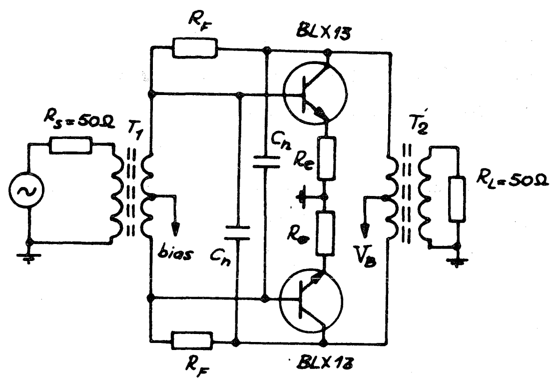


Fig. 1

The minimum required D.C. input power to the transistors can be derived from the facts that:

a. A single transistor is able to deliver 8 Watts P.E.P. at -40dB I.M.D. and at 28 Watts D.C. input power.

b. the power required to drive the tube is appr. 12 Watts P.E.P. also at -40dB I.M.D.

Therefore the D.C. input power to the transistors must be at least:
12/8 x 28 = 42 Watts.

3.2 The amplifier

Fig. 2 shows the simplified H.F. circuit (one side) in which V_{CE} and I_c are in phase.

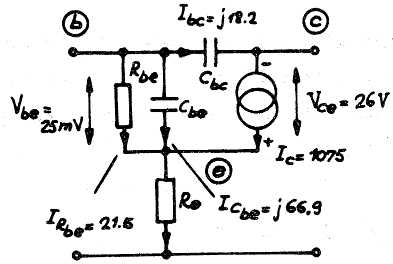


Fig. 2

To make an output transformer superfluous the ratio V_{ce}/I_c must be appr. 50 Ohms. However the transformer is needed to realize an asymmetrical output from the push-pull set-up.

As Fig. 2 shows an emitter resistor Re has been applied. The voltage drop is appr. 2 Volts so V_{ce} = 26 Volts has been chosen for V_B = 28 Volts. For

T_{hs} = 70°C a D.C. dissipation of 28 Watts per transistor is allowed.

I_c = 28/26 = 1.075 A and the max. P_o is 50% of 28 Watts or 14 Watts per transistor.

This operating point related to the f_T curve is acceptable.

Because the max. P_o is at least 20% less due to losses in feedback resistors an output of 22 Watts (from 2 transistors) for IMD = -30dB or appr. 15 Watts for -40dB can be expected.

$I_{C_{be}}$ can be calculated from the relation $f_T/f_w = I_c/I_{C_{be}}$.
With f_T typ. = 450 MHz and f_w = 28 MHz $I_{C_{be}} = I_c . f_w/f_T$ = 1.075x28/450 = 66.9

Calculating with h_{FE} = 50, $I_{R_{be}}$ becomes 1.075/50 = 21.5 mA.

$I_{R_{be}}$ and I_c add in R_e (Fig. 3)

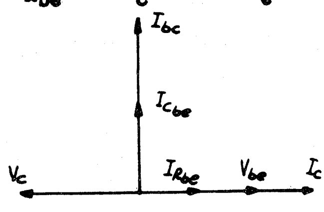


Fig. 3

Hence: I_{R_e} = 1075 + 21.5 = appr. 1096 mA

Because $V_{be} = k.T/q \approx 25$ mV at 25°C:
V_{bc} = 26.025 Volts.

The typical value of C_{bc} at V_{CE} = 26 Volts is appr. 28 pF.
Due to A.C. voltage excitation C_{bc} will rise by 10% to appr. 31 pF.

To prevent oscillation by overneutrodynisation C_n = 27 pF has been chosen. It can be demonstrated that C_n may be subtracted from the present C_{bc}, so:
C'_{bc} = 31 - 27 = 4 pF, what is equal to -j1.43 kOhms at 28 MHz.
All values have been calculated at the upper frequency so

$$I_{C_{bc}} = 26.025/-j\ 1.43 \times 10^3 = j\ 18.2 \text{ mA.}$$

The transistor (one side) can now be inserted in the total feedback chain, whilst by means of T_1 the input is adapted to 50 Ohms. Fig. 4 shows the extended circuit.

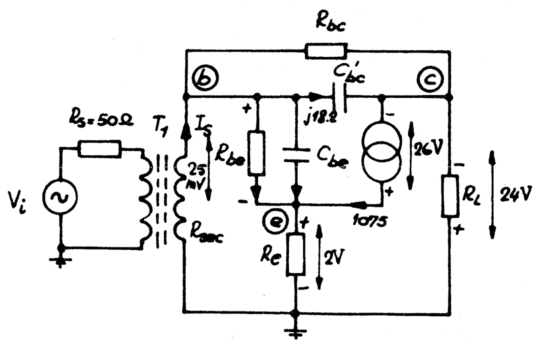


Fig. 4

For $I_{R_{bc}}$ it holds that:

$$I_{R_{bc}} \geqslant I_{C_{bc}} + I_{C_{be}} - I_{R_{be}} = 63.6 \text{ mA}$$

To avoid a compensation inductance in series with R_{bc} it is even desirable to choose $I_{R_{bc}}$ 1.5 to 2 times the above mentioned value.
On the other hand this current must not be chosen too high because it reduces the available output current. A good compromise is therefore 10% of I_c or 107.5 mA.
The feedback resistor $R_{bc} = (V_{ce} + V_{be})/I_{R_{bc}} = (26.025)/107.5 \times 10^{-3} = 242$ Ohms
and $R_e = V_{R_e} / (I_{R_{be}} + I_c) = 2/\ (21.5 + 1075) = 1.82$ Ohm.

Because of the internal $r_e \approx 0.2$ Ohm the external R_e has to be $\approx$1.62 Ohm.

The source current $I_s = I_{R_{bc}} + I_{R_{be}} = 107.5 + 21.5 = 129$ mA.

Hence: $R_{sec} = (V_{R_e} + V_{R_{be}})/I_s = (2.025)/129 \times 10^{-3} = 15.7$ Ohms.

$P_{i_{eff}} = I_s\,(V_{R_e} + V_{R_{be}})/2 = 129 \times 10^{-3} \times 2.025/2 = 130.5$ mWatts

With reference to Fig. 4 it can be seen that the output peak voltage is:

$V_{ce} - V_{R_e} = 26 - 2 = 24$ Volts and

$I_{R_L} = 1075 - 107.5 = 967.5$ mA.

$R_L = (V_{ce} - V_{R_e})/I_{R_L} = 24/\,0.9675 = 24.8$ Ohms

$P_{o_{eff}} = (V_{ce} - V_{R_e})^L\, I_{R_L}/2 = 24/2 \times 0.9675 = 11.6$ Watts.

The power gain $A = 10 \log (P_{o_{eff}}/P_{i_{eff}}) = 10 \log (11.6/0.1305) = 19.5$dB

3.2 The input transformer

Wide-band transformers for the h.f. band can be executed as conventional and as transmission line type.

Extensive information on both types one will find in CAB reports NCO 6814 [3], ECO 6907 [4], ECO 7113 [2] and ECO 7114 [1].

In order to prevent difficulties with the set-up of a wide-band transmission line type with the ratio $T^2 = (n_{prim}/n_{sec})^2 = 1.59$ a conventional type with separate windings has been applied.

The reactance of the shunting inductance at the lowest frequency 1.6 MHz has been chosen to be at least 4 times the primary (source) resistance of 50 Ohms, what roughly corresponds with an input VSWR of $1 + R_s/X_L = 1.25$, a value being improved by compensation later on.

So the inductance is 20 μH or +j200 Ohms at 1.6 MHz.

The transformer is wound on a ferrite toroid of 4C6 material.

Earlier investigations showed that this is a good material for this frequency range. These toroids are delivered in standard dimensions and it must first be investigated which size is needed.

This can be done with the aid of the formulae:

$$L = \mu_o . \mu_r . n^2 . A/l \qquad \text{and} \qquad B_{max.} = V_{max}/\omega . A.n.$$

In these equations A is the average ferrite cross section in m^2 and l the average length of the lines of force in m.
Both A and l are unknown.

Finding the number of turns n from the first equation and introducing it in the second one gives after rearranging

$$A.l = (V_{max}/\omega . B_{max})^2 . \mu_o . \mu_r / L$$

in which:

$$\mu_o = 4.\pi . 10^{-7}$$
$$\mu_r = 120 \pm 20\% \quad (4C6)$$
$$L = 20 \mu H$$
$$\omega = 2 \pi \times 1.6 \times 10^6 \text{ rad/sec.}$$

V_{max} can be calculated from the power that has to be transferred.

As calculated before the input power in the push-pull configuration amounts to appr. 2 x 130.5 = 261 mWatts.
Taking into account eventualities (VSWR, lower gain, etc.) it has been increased to P_i = 500 mWatts.

In that case: $V_{max} = (2. P_i . R_s)^{\frac{1}{2}} = 2 \times 0.5 \times 50^{\frac{1}{2}} = 7.07$ Volts.
B_{max} depends on the parallel loss resistance that is acceptable what can be calculated from Figs. 6 and 7.

Fig. 6 shows $\mu_r R_p/L$ versus $B_{max}.f$ for 1.6 MHz and Fig. 7 $\mu_r R_p/L$ versus frequency f for $B_{max} \rightarrow 0$.

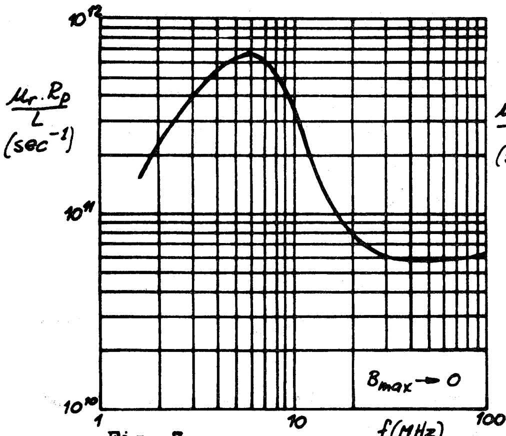


Fig. 7

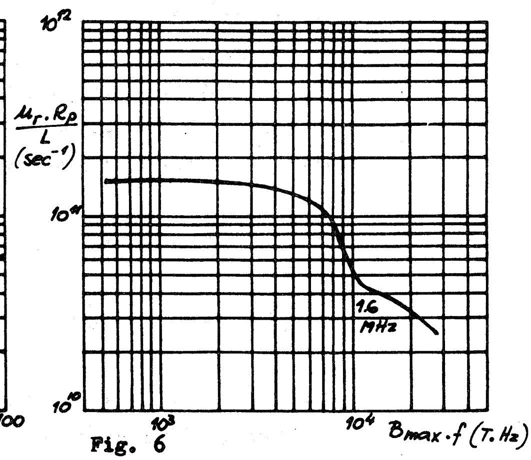


Fig. 6

If a power loss of 1% is acceptable then $R_p = 100 \times 50 = 5$ kOhms. Hence $\mu_r R_p/L = 120 \times 5 \times 10^3/20 \times 10^{-6} = 3 \times 10^{10}$ sec^{-1}. As at high frequencies R_p depends hardly on B_{max}, we may conclude from Figs. 6 and 7, that above calculated value can only occur at 1.6 MHz.

The value of $B_{max}.f$ corresponding with above mentioned $\mu_r R_p/L$ amounts to 2×10^4 T.Hz according to Fig. 6, so the resulting value for $B_{max} = 2 \times 10^4/1.6 \times 10^6 = 1.25 \times 10^{-2}$ T at 1.6 MHz.

The required product A.l can now be determined:

$$A.l = (7.07/2\pi \times 1.6 \times 10^6 \times 1.25 \times 10^{-2})^2 \times 4\pi \times 10^{-7} \times 120/20 \times 10^{-6} = 0.0241 .\ 10^{-6}\ m^3$$

In the table below the A.l products of some available toroids have been given, calculated according to:

$$A.l = \left\{h\ (D - d)/2\right\} \left\{\pi (D + d)/2\right\}$$

Catalog number	Dimension (mm) D x d x h	A (m^2)	l(m)	A.l (m^3)
4322 020 91000	6 x 4 x 2	2×10^{-6}	1.55×10^{-2}	0.0310×10^{-6}
4322 020 91010	9 x 6 x 3	4.51×10^{-6}	2.33×10^{-2}	0.1050×10^{-6}
4322 020 91020	14 x 9 x 5	12.54×10^{-6}	3.55×10^{-2}	0.445×10^{-6}
4322 020 91070	23 x 14 x 7	31.4×10^{-6}	5.70×10^{-2}	1.790×10^{-6}

As the table shows the smallest toroid that meets our needs is a type with the dimensions 6 x 4 x 2 mm.
The number of turns for an inductance of 20 μH is calculated with the aid of the rearranged equation:

$$n = (L.l/\mu_o.\mu_r.A)^{\frac{1}{2}} = (20x10^{-6}x1.55x10^{-2}/4\pi x10^{-7}x120x2x10^{-6})^{\frac{1}{2}} = 32.1$$

Together with the secondary (2 x 13 turns in this case) this results in 58 turns on a toroid of rather small dimensions.
A more practical size is the 14 x 9 x 5 mm toroid.

In that case the number of turns becomes:

$$n = (20 \times 10^{-6} \times 3.55 \times 10^{-2}/4\pi \times 10^{-7} \times 120 \times 12.5 \times 10^{-6})^{\frac{1}{2}} = 19.4$$

rounded to 20 turns $\rightarrow$ L = 21 μH.

The number of secundary turns depends on the impedance step to be made:
$n_{sec} = n_{prim}\ (R_p/R_s)^{-\frac{1}{2}} = 20\ (50/31.4)^{-\frac{1}{2}} = 16.4$; because of mid-tap supply a 2 x 8 turns secundary has been chosen.

The maximum wire diameter d' can be obtained from:

$$d' = \pi d/(n_{tot} + \pi) = \pi x\ 9/(36 + 3.14) = 0.717;$$

a thickness of 0,6 mm CuEm wire has been chosen.

Due to the higher A.l product of the coil the B_{max} has been changed to:

$$B_{max} = V_{max}/\omega.A.n = 7.07/2\pi x1.6x10^6 \times 12.54x10^{-6}x20 = 2.81 \times 10^{-3}\ T$$

The corresponding value for $\mu_r\ R_p/L$ at 1.6 MHz is 1.3×10^{11} ($B_{max}\ f = 4.5 \times 10^3$). The value of R_p is

$$R_p = L\ x1.3 \times 10^{11}/\mu_r = 21 \times 10^{-6} \times 1.3 \times 10^{11}/120 = 22.8\ \text{kOhms.}$$

So the core loss at 1.6 MHz is $R_L/R_p = 50/22.8 \times 10^3 = 2.2\ ^o/oo$

3.3 The output transformer (50 Ohms design)

The output transformer has the same lay-out as the input transformer and the calculation of it is more or less similar to that of the input type.

The volume of the ferrite core of 4C6 material is calculated again by $A.l = (V_{max}/\omega . B_{max})^2 . \mu_o . \mu_r / L$ in which μ_o, μ_r, L, ω and B_{max} are equal to those for the input transformer.

V_{max} can be calculated from the max. output power and the impedance R_L of 50 Ohms.
If this case it has been assumed that the maximum power can amount to 25 Watts, what is equal to $V_{max} = (2.\ P_o .\ R_L)^{\frac{1}{2}} = (2 \times 25 \times 50)^{\frac{1}{2}} = 50$ Volts over 50 Ohms.

$$A.l = (50/2\pi \times 1.6\times10^6 \times 1.25\times10^{-2})^2 \times 4\pi\times10^{-7}\times120/20\times10^{-6} = 1.14\times10^{-6}\ m^3$$

The table in chapter 3.2 shows that it is obvious to take the 23 x 14 x 7 mm toroid core, having an A.l product of $1.79 \times 10^{-6}\ m^3$.

The number of turns for a 20 μH secundary follows from:

$$n = (L.l/\mu_o . \mu_r . A)^{\frac{1}{2}} = (20 \times 10^{-6} \times 5.82 \times 10^{-2} / 4\pi\, 10^{-7} \times 120 \times 31.4 \times 10^{-6})^{\frac{1}{2}} = 15.65$$

rounded to 16 turns $\longrightarrow$ L = 21 μH

Primary, the impedance is 2 x 24.8 = 49.6 Ohms, what determines the number of turns to: $n_{prim} = n_{sec}\ (R_s/R_p)^{-\frac{1}{2}} = 16\ (50/49.6)^{-\frac{1}{2}} = 15.95$;
because of mid-tap supply 2 x 8 turns should be chosen. However experiences later on showed better performance for 2 x 7 turns primary. So further calculations and measurements shall be made with that ratio.

The max. wire diameter d' is: 3.14 x 14/(30 + 3.14) = 1.31 mm, a thickness of 1.0 mm $C_u E_m$ wire has been taken.

In the first instance B_{max} has been assumed to be a constant, however R_p can vary by increasing the core size.

$$B_{max} = V_{max}/\omega.A.n = 50/2 \times \pi \times 1.6\times10^6 \times 31.4\times10^{-6} \times 16 = 9.88 \times 10^{-3}\ T.$$

The corresponding value for $\mu_r R_p/L$ at 1.6 MHz is 3.5×10^{10}, ($B_{max}.f = 1.58 \times 10^4$).

$R_p = L\times3.5\times10^{10}/\mu_r = 20\times10^{-6}\times3.5\times10^{10}/120 = 5.83$ kOhms, what means a core loss at 1.6 MHz of $R_L/R_p = 50/5.83 \times 10^3 = 0.86\%$.

3.4 Compensation

3.4.1 Spreading inductance, uncompensated transfer

To prevent an unacceptable spreading in both designs the primary and secondary windings have been distributed over the whole circumference. Consequently the primary has been wound in between the secondary. In that case it is advisable to put each corresponding primary and secondary winding close together.

With the aid of a Vector Imp. meter (HP 4815A) the behaviour of the uncompensated input- and output transformers has been measured.
As the tables show there is a deviation in the highest octave.
Measurements were resp. made with an ohmic termination of: 31.4 Ohms secondary for the input trafo (Fig. 8a) and 38.3 Ohms primary for the output transformer (Fig. 8b).

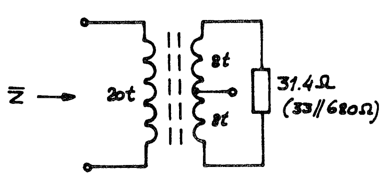


Fig. 8a Input trafo

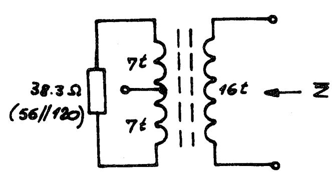


Fig. 8b Output trafo

	Input trafo		Output trafo	
f (MHz)	$\bar{Z}$ (Ohms)	VSWR	$\bar{Z}$ (Ohms)	VSWR
1.6	47.3	1.26	46.8	1.32
3.5	48.2	1.19	49.0	1.21
7.0	49.3	1.22	49.1	1.24
14	49.7	1.40	50.3	1.43
20	51.2	1.58	50.7	1.61
28	53.6	1.89	53.0	1.93

The spreading inductance is:

f (MHz)	X_L (Ohms)	L (nH)	X_L (Ohms)	L (nH)
1.6	1.96	195	2.04	203
28	30.78	175	33.89	193

3.5 H.F. compensation of the input transformer

Fig. 9 shows the applied system seen from the primary.

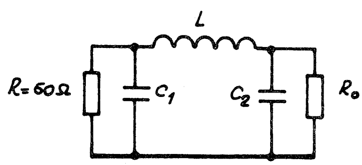


Fig. 9

R_o is the transformed output resistance and L the spreading inductance seen from the primary.

According to Nielinger $\boxed{5}$ two capacitors $C_1 = C_2$ will be added to form an L.P. section by which the max. VSWR is reduced.

C_2 is transformed from the secondary and consists partly of the transistor input capacitances.

$X_{L\,max} = 2\pi \cdot f_{max} \cdot L = 30.78$ Ohms

$f_{max} = 28$ MHz

$X_{L\,max}/R = 30.78/50 = 0.616$ $\therefore$ VSWR max $= 1.016$

From the curve in $\boxed{5}$ it follows that:

$R/X_{C_{min}} = 0.33$ $X_{C_{min}} = 50/0.33 = 152$ Ohms.

$$X_{c_{min}} = 1/\ 2\pi .\ f_{max}.\ C$$
$$C = 1/\ 2\pi \times 28 \times 10^6 \times 152 = 37.3\ pF$$

In fact the latter is the primary capacitance, whilst the secondary becomes:
$T^2 .\ C = 1.59 \times 37.3 = 59.4\ pF.$

The primary capacitance has been chosen 2 x 18 = 36 pF, whilst the secondary one has been determined experimentally because of the input capacitance of the transistors. A value of 10 pF for this capacitance gives the smallest gain variation and the lowest input VSWR through the band.

3.6 L.F. compensation of the input transformer

Fig. 10 shows the essential elements for a simple high-pass section at 1.6 MHz.

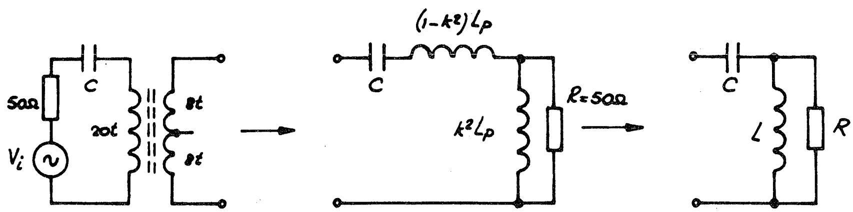


Fig. 10a Fig. 10b Fig. 10c

Primary inductance at f = 1.6 MHz is 21 μH (j 211 Ohms).
According to $\boxed{5}$:

$$R/X_{L_{min.}} = X_{c_{max}}/R$$
$$X_{L_{min}} = 211\ \text{Ohms}$$
$$R/X_{L_{min}} = 50/211 = 0.237 \quad \therefore\ \text{VSWR} = 1.06$$
$$X_{c_{max}} = 0.237 \times 50 = 11.85$$
$$C = 1/\ 2\pi .f_{min}.X_{c_{max}} = 1/\ 2\pi \times 1.6 \times 10^6 \times 11.85 = 8.4\ nF;$$ 8.2 nF has been chosen.

3.7 The compensated input transformer

The compensated transformer (Fig. 11) has been checked with a vector impedance meter. First the parasitic properties of the mount have been determined. The results have been put in a small digital computer, which output shows the performance versus f given in R_s, X_s and VSWR. This method will be applied in subsequent measurements.

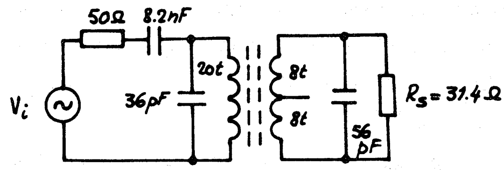


Fig. 11

f (MHz)	R_s (Ohms)	X_s (Ohms)	VSWR
1.6	48.0	0.68	1.04
3.5	49.2	0.51	1.02
7.0	49.9	0.18	1.00
14	50.9	0.91	1.03
20	52.5	-0.44	1.05
28	55.9	1.34	1.12

3.8 H.F. compensation of the output transformer

The calculation is in main lines similar to the foregoing, so it shall be given briefly.

38.3 → 50 Ohms design

$X_{L_{max}}$ = $2\pi \cdot f_{max}$ = 33.9 Ohms

f_{max} = 28 MHz

$X_{L_{max}}/R$ = 33.9/50 = 0.678 ∴ $VSWR_{max}$ = 1.025

From the curve in $\boxed{5}$ it follows that:

$R/X_{c_{min}}$ = 0.37 $X_{c_{min}}$ = 50/0.37 = 145

$X_{c_{min}}$ = $1/2\pi \cdot f_{max} \cdot C$

C = $1/2\pi \times 28 \times 10^6 \times 145$ = 39 pF (across the secondary)

Because T^2 = 1.31 (2 x 7 → 16 turns), the capacitance across the primary must be 51 pF, because of the transistor output and neutralizing capacitances a value of 22 pF has been chosen.

3.9 L.F. compensation of the output transformer

Because L = 21 µH, the calculation is similar to that of the input transformer, so the coupling capacitor is 8.2 nF.

3.10 The compensated output transformer

The complete transformer design (Fig. 12) has been checked with the vector impedance meter. The most near-by standard capacitor value has been taken.

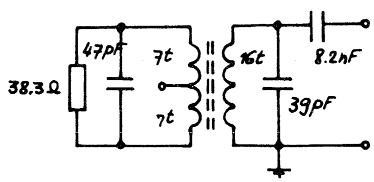


Fig. 12

f (MHz)	R_s (Ohms)	X_s (Ohms)	VSWR
1.6	48.5	0.69	1.03
3.5	50.0	0.53	1.01
7.0	50.4	0.19	1.01
14	51.9	0.96	1.04
20	52.9	0.97	1.06
28	56.3	2.37	1.13

3.11 D.C. equivalent circuit

Fig. 13 shows the circuit.

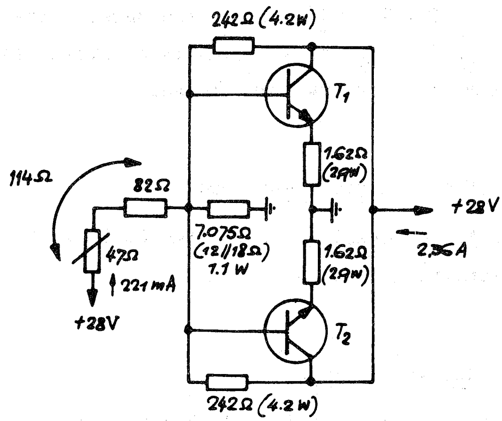


Fig. 13

The 47 Ohms pot. meter has to be adjusted to I_T = 2.58 A for a supply voltage of 28 Volts. As the parts list shows, the feedback resistors have been composed of several ones in parallel. Particularly the emitter resistors need to have small series-inductances.

4. Measured results

Measurements have been carried out under the following conditions unless stated otherwise:

V_B = 28 Volts
R_s, R_o = 50 Ohms
Ambient and stud temperature $T_A = T_S$ = 25°C.

4.1 Intermodulation distortion

The I.M.D. versus output power is measured with a two tone signal (p, q) at the spot frequencies 1.6, 3.5, 7.0, 14, 20 and 28 MHz.
This signal is made as follows: The output signals of two X-tal oscillators with a difference of 1 kHz, switchable in pairs at above frequencies, have been separately amplified in two linear wide-band amplifiers. These amplified signals have been combined in a hybrid and via switchable low-pass filters supplied to the amplifier under test.

The oscillator kit also contains X-tal oscillators presenting the local oscillator signals to the spectrum analyzer SINGER MF5. Fig. 14 shows the block diagram.

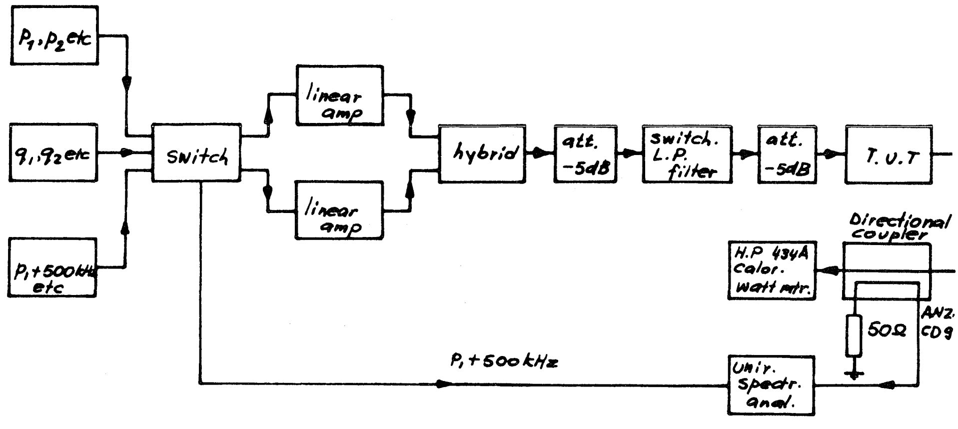
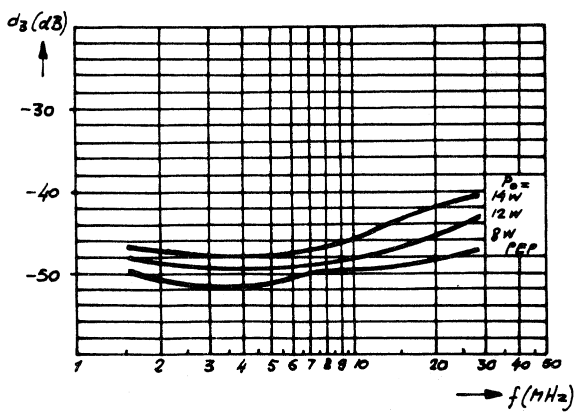


Fig. 14

In Fig. 15 the IMD, expressed in d_3, is given versus frequency.
If the suppression of the d_5 product is less than -60dB the measured value has been given in the graph (Fig. 15).



Except for the point f = 28 MHz, P_o = 14 Watts PEP where the d_5= -58dB the d_5 is better than -60dB.

Fig. 15

4.2 Input VSWR and power gain

These values have been measured with the set-up of Fig. 16.

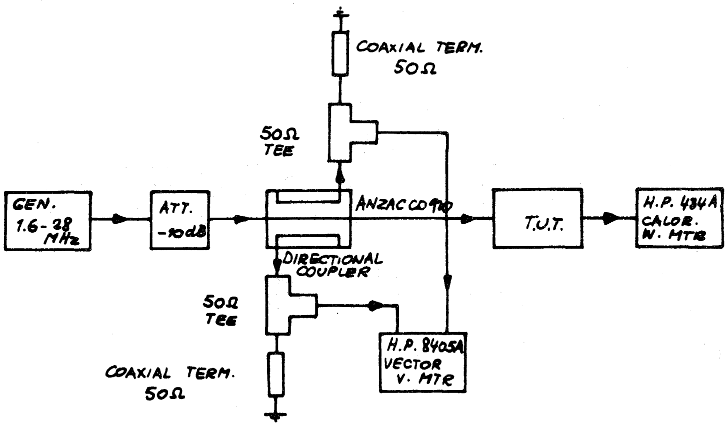


Fig. 16

The measurements have been carried out for P_o = 1 Watt (Fig. 17).

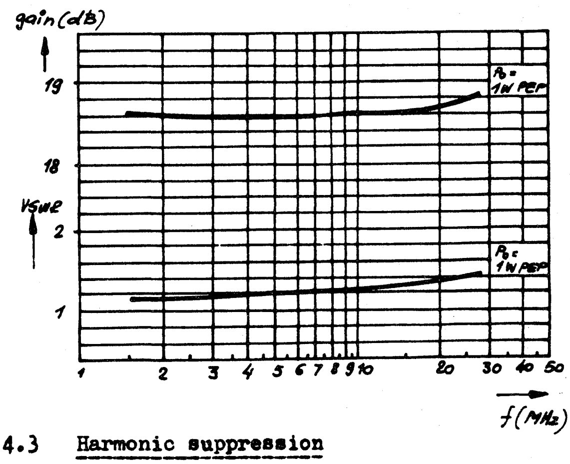


Fig. 17

4.3 Harmonic suppression

The in-band harmonic content, the 2nd and 3rd harmonics 2p and 3p, has been measured with the set-up of Fig. 18.
Measurements have been done with single tone signals (p) of 1.6, 3.5, 7.0, 14, 20 and 28 MHz, being available from the two-tone equipment.

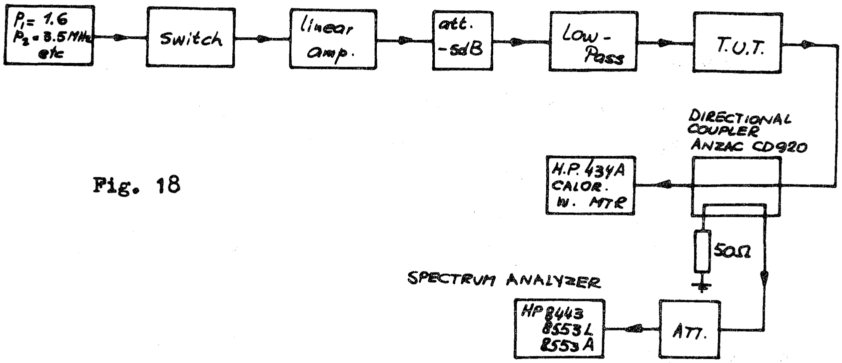


Fig. 18

P_o = 14 Watts

f :	1.6 MHz	3.5 MHz	7.0 MHz	14 MHz	20 MHz	28 MHz
2nd harmonic (2p):	-34dB	-33dB	-28dB	-38dB	-43dB	-39dB
3rd harmonic (3p):	-14dB	-14dB	-21dB	-26dB	-29dB	-30dB

Because the results could be influenced by variation in sensitivity of the coupler for higher frequencies, the latter has been checked. It appeared that the measuring results were correct.

5. Extension to tube driver

5.1 The input of the YL1230

As mentioned before the final stage to be driven by this amplifier is the air cooled RF power tetrode YL1230. The tube can supply in class AB (V_A = 3 kV) 1050 Watts PEP output in the load for a peak driving voltage Vg_1 p = 46 Volts (< 51 Volts) up to 30 MHz.
Under these operating conditions the IMD has been given as:

1 MHZ: $d_3 < -38dB$; $d_5 < -38dB$.
30 MHz: $d_3 < -36dB$; $d_5 < -36dB$.

The average input capacitance of the tube is equal to:

$$C_i = Cg_1 - k,f + Cg_1g_2 + (A_v + 1)\ Cag_1 + C_{wiring} =$$

$$37.5 + 56 + (50).\ 0.1 + 3 = 101\ pF$$

because:

$Cg_1 - k,f$ = $33 \div 42$ pF avg. 37.5 pF
Cg_1g_2 = $48 \div 64$ pF avg. 56 pF
Cag_1 = < 0.1 pF
C_{wiring} = ~3 pF
Voltage gain A_v is appr. 50x (V_A = 3 kV, Vg_2=560 V, $Vg_{1\ p}$= 46V, Va_p=2200V)

5.2 The low-pass filter

To drive the YL1230 to its full output power a drive of 46 Volts peak has to be developed across the input capacitance of 101 pF.
Because the tube, operating in class-AB_1, does not draw grid current the parallel damping is very small.

A similar case has been solved about 2 years ago, when 2 pieces 2N3632 had to drive a YL1070 or a YL1150. For that purpose the low-pass filter configuration, shown in Fig. 19, was used. In this filter C_k represents the tube input capacitance.
Starting from a known Chebyshev design, the inductances L_k were varied until a good compromise was reached between gain flatness and input VSWR.

Later a computer-program has been written for this purpose in which all components could be varied individually; even a mutual inductance between the coils was considered. The best compromise is given in general form (see Fig. 20).

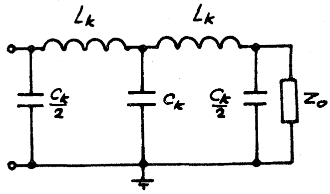


Fig. 19

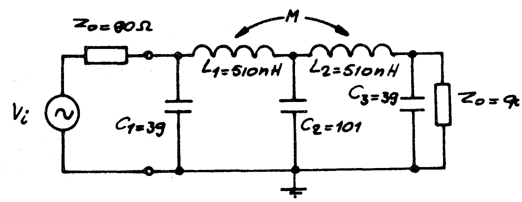


Fig. 20

$Z_o = 9.09 \times 10^{-9}/C_2$

$C_1 = C_3 = 0.386\ C_2$

$L_1 = L_2 = 0.515 \times 10^{-16}/C_2$

$M = 0$

Maximum VSWR: 1.36

Variation of V_{c_2}: ± 0.363dB

Maximum required input voltage (on C_1): $0.937\ V_{c_2}$

Maximum required input current: $1.17\ V_{c_2}/Z_o$

Maximum required input power : $0.575 \times 10^8\ C_2\ V_{c_2}^2$.

In our specific case the element values become as indicated in Fig. 20. The table below gives the response, the input VSWR and the required input voltage and current versus frequency.

f (MHz)	A (dB)	VSWR	V_{in} (V)	I_{in} (A)
1.6	-0.006	1.049	45.89	0.5133
2.5	-0.014	1.077	45.79	0.5163
3.5	-0.027	1.108	45.50	0.5211
5.0	-0.054	1.154	45.00	0.5307
7.0	-0.099	1.212	44.14	0.5466
10	-0.174	1.285	42.65	0.5725
14	-0.252	1.344	40.92	0.5991
20	-0.194	1.345	40.25	0.5984
24	+0.043	1.326	40.35	0.5698
28	+0.474	1.357	40.30	0.5279

5.4 The modified transformer (38.3 → 90 Ohms)

Maintaining the primary number of turns at 2x7, the secondary becomes 21 turns for the same core size 23 x 14 x 7 mm.
The secondary inductance has been obtained from the equation:

$$L = n^2 \cdot \mu_r \cdot \mu_o \cdot A/l = 21^2 x 4\pi \times 10^{-7} x 120 x 31.5 x 10^{-6}/5.82 x 10^{-2} = 36.1\ \mu H$$

The spreading inductance has been measured:

f (MHz)	X_L (Ohms)	L (nH)
1.6	3.46	344
28	60.36	345

The max. wire diameter for n_{tot} = 35 turns becomes: 3.14 x 14/(35+3.14) = 1.16 mm.
1.0 mm CuEm has been applied.

5.5 H.F. compensation of the 90 Ohms output transformer

Calculation is similar to the one described in chapter 3.8.

$X_{L_{max}} = 2\pi \cdot f_{max} = 60.4$

$f_{max} = 28$ MHz

$X_{L_{max}}/R = 60.4/90 = 0.672 \quad \therefore$ VSWR max = 1.025

From the curve in $\boxed{5}$ it follows that:

$R/X_{c\ min} = 0.37 \qquad X_{c\ min} = 90/0.37 = 243$

$X_{c_{min}} = 1/2\pi . f_{max} . C$

$C = 1/2\pi \times 28 \times 10^6 \times 243 = 23.3$ pF (across the secondary)

Because $T^2 = 2.25$ (2x7 $\rightarrow$ 21 turns), the capacitance across the primary must be 52.5 pF.
Because of the transistor output and neutralizing capacitances a value of 22 pF has been chosen.

5.6 L.F. compensation of the 90 Ohms output transformer

Secondary inductance at f = 1.6 MHz is 36.1 µH (j 362.5 Ohms).
According to $\boxed{5}$:

$R/X_{L\ min} = X_{c_{max}}/R$

$X_{L_{min}} = 362.5$ Ohms

$R/X_{L_{min}} = 90/362.5 = 0.248 \quad \therefore$ VSWR = 1.01

$X_{c\ max} = 0.248 \times 90 = 22.3$

$$C = 1/2\pi \cdot f_{min} \cdot X_{c\ max} = 1/2\ \pi \times 1.6 \times 10^6 \times 22.3 = 4.45\ nF$$

A value of 5.6 nF has been chosen to reduce the load impedance of the transistors at low frequencies slightly, which improves the I.M.D. somewhat.

5.7 The compensated output circuit

Fig. 21 shows the combination of the networks.

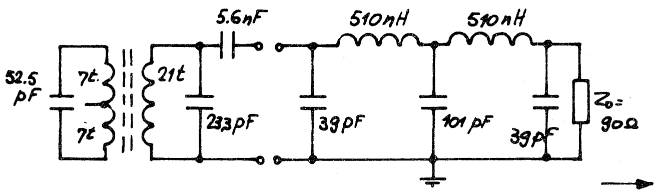


Fig. 21a

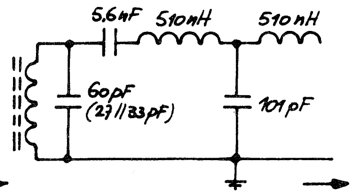


Fig. 21b

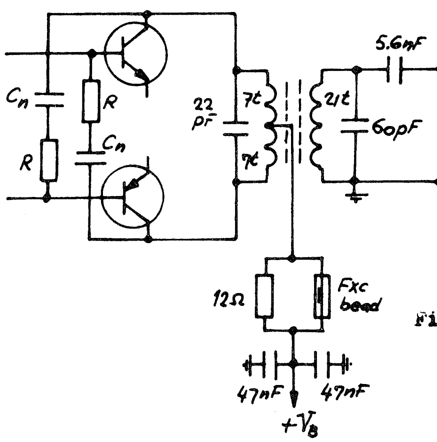


Fig. 21c

The secondary capacitance has been composed of 27 + 33 = 60 pF, whilst the primary one is 22 pF because of already present output- and neutrodyne capacitances of the BLX13's.

To prevent parasitic oscillations in the VHF range a resistor of 33 Ohms has been used in series with each neutrodyne capacitor. This combination represents an increasing damping to the transistor versus frequency.

The decoupling method for the supply voltage (Fig. 21c) has been applied because it functions like a hybrid; the same has been done at the input side.

6. Measured results (tube driver)

6.1 Intermodulation distortion

In Fig. 22 the IMD, expressed in d_3, is given versus frequency.
The parameter is the output voltage across the 101 pF capacitor, replacing the input capacitance of the YL1230.
Set up: Modified circuit of Fig. 14 (see Fig. 23).

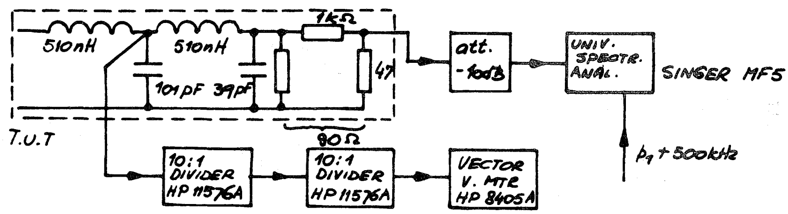
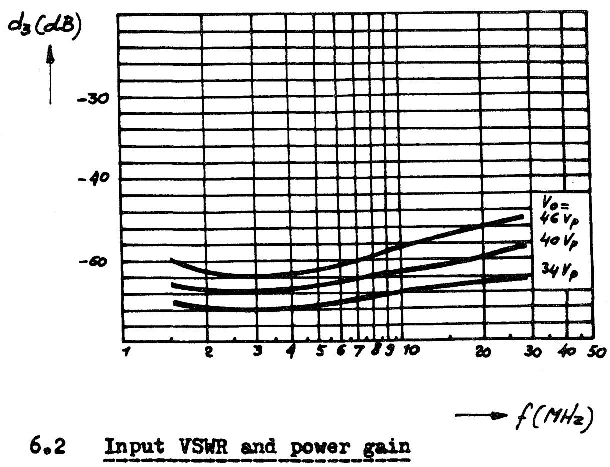


Fig. 23

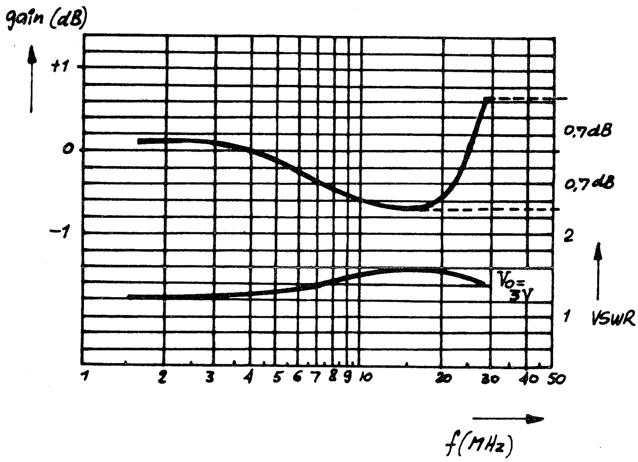


d_5 is better than -60dB

Fig. 22

6.2 Input VSWR and power gain

Set-up: modified circuit of Fig. 16. The output voltage has been measured in the way shown in Fig. 23.



The drive power
P_{dr} = 142 mWatts
± 0.7dB

OdB level: 142 mW → $46V_I$

Fig. 24

6.3 Harmonic suppression

Set up: Modified circuit of Fig. 18 (see Fig. 24a)

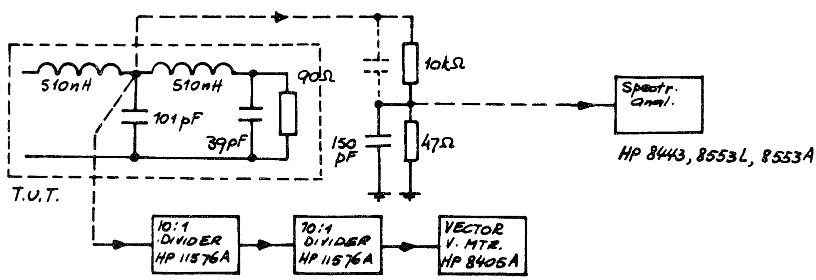


Fig. 24a

$\hat{V}_o$ = 46 Volts

f	1.6 MHz	3.5 MHz	7.0 MHz	14 MHz	20 MHz	28 MHz
2nd harmonic (2p):	-46dB	-49dB	-49dB	-48dB	-45dB	-42dB
3rd harmonic (3p):	-42dB	-40dB	-41dB	-37dB	-40dB	-60dB

7. Schematic diagram and mechanical lay-out

Figs. 25 and 26 resp. show the 50-50 Ohms amplifier and the filter unit, whilst the drawing of the p.c. board with situation of the components has been given in Figs. 27a, b, c, d.

The lower sheet of the double clad p.c. board functions as a ground plane. Interconnections of some upper parts with the ground plane were made with 2 mm tubular rivets. These rivets were soldered to the print conductors to be sure of contact.

Because the amplifiers contain emitter resistors a part of the ground plane has been insulated.

During measurements the board is placed on a water-cooled heatsink.

8. References

1. J. Mulder: "A single-stage wide band linear power amplifier for 80-100 W PEP in the 1.6-28 MHz frequency band equipped with two pieces BLX14". CAB report ECO 7114.

2. M.J. Köppen: "Single stage wide-band (1.6-28 MHz) SSB driver modules with BLY92A and BLX13, operating in class A". CAB report ECO 7113.

3. J.M. Siemensma: "Investigations on linear power amplifiers for SSB wide-band class A operation from 1.6 to 28 MHz". CAB report NCO 6814.

4. A.H. Hilbers: "On the design of H.F. wide band power transformers". CAB report ECO 6907.
or: "Design of H.F. wide band power transformers". Application Information 530.

5. H. Nielinger: "Optimale Dimensionierung von Breitbandanpassungs-netzwerken". NTZ Heft 2 1968 pp 88-91.

6. J. Mulder: "A Linear Solid-State Wideband Driver for Medium Power S.S.B. Tubes (f = 1.6 - 28 MHz)". CAB report ECO 6917.

M.J. Köppen

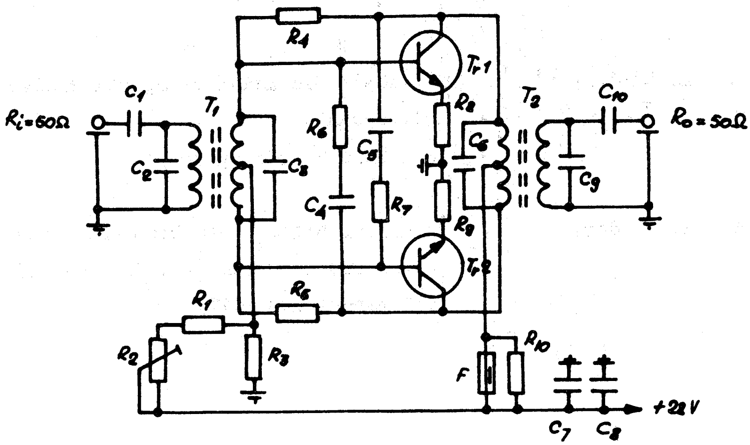

R4
Tr1
C1
T1
R8
T2
C10
Ri=60Ω
Ro=50Ω
R6
C5
C6
C2
C3
C9
C4
R7
R9
Tr2
R1
R5
R10
R2
R3
F
+22V
C7
C8

Fig. 25

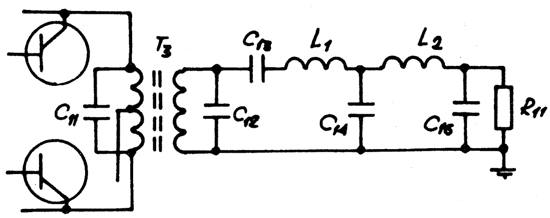

T3
C13
L1
L2
C11
C12
C14
C16
R11

Fig. 26

9. **Components**

R_1	=	82 Ohms, enamelled wire-wound 5.5 Watts	2322 320 31479
R_2	=	47 Ohms wire-wound trimming pot. meter, 2 Watts	2322 011 02479
R_3	=	parallel connection of:	
		1x 12 Ohms, carbon ±5% CR52 style	2322 101 63129
	and	1x 18 Ohms, carbon ±5% CR52 style	2322 101 63189
R_4, R_5	=	parallel connection of:	
		4x 1 kOhms, carbon ±5%, CR68 style	2322 214 13102
	and	1x 27 kOhms, carbon ±5%, CR37 style	2322 212 13273
R_6, R_7	=	33 Ohms, carbon ±5%, CR37 style	2322 212 13339
R_8, R_9	=	parallel connection of:	
		6x 10 Ohms, carbon ±5%, CR37 style	2322 212 13109
R_{10}	=	12 Ohms, carbon ±5%, CR37 style	2322 212 13129
R_{11}	=	parallel connection of:	
		11 x 1 kOhms, carbon ±5%, CR52 style	2322 101 63102
	and	1x 8.2 kOhms, carbon ±5%, CR52 style	2322 101 63822
C_1	=	8.2 nF polyester ±10%	2222 342 45823
C_2	=	parallel connection of:	
		2x 18 pF ceramic	2222 555 56189
C_3	=	10 pF ceramic	2222 555 56109
C_4, C_5	=	27 pF ceramic	2222 555 56279
C_6	=	22 pF ceramic	2222 555 56229
C_7, C_8	=	47 nF polyester ±10%	2222 342 45473
C_9	=	39 pF ceramic	2222 555 56399
C_{10}	=	8.2 nF polyester ±10%	2222 342 45823

C_{11}	=	22 pF ceramic	2222 555 56229
C_{12}	=	parallel connection of:	
		1x 27 pF ceramic	2222 555 56279
	and	1x 33 pF ceramic	2222 555 56339
C_{13}	=	5.6 nF polyester ±10%	2222 342 45563
C_{14}	=	input capacitance of YL1230 (101 pF)	
C_{15}	=	39 pF ceramic	2222 555 56399
F	=	ferroxcube bead grade 3B	4312 020 31500

$L_1 = L_2$ = 510 nH, $7^3/4$ turns of CuEm wire, close wound on D = 10 mm.

T_1 = 4C6 ferroxcube toroid 14 x 9 x 5 mm

prim.: 20 turns 0.6 mm CuEm wire
sec. : 2 x 8 turns 0.6 mm CuEm wire (see Fig. 27a)

T_2 = 4C6 ferroxcube toroid 23 x 14 x 7 mm

prim.: 2 x 7 turns 1.0 mm CuEm wire
sec. : 16 turns 1.0mm CuEm wire (see Fig. 27a)

T_3 = 4C6 ferroxcube toroid 23 x 14 x 7 mm

prim.: 2 x 7 turns 1.0 mm CuEm wire
sec. : 21 turns 1.0mm CuEm wire (see Fig. 27d)

$Tr_1 = Tr_2$ = BLX13.

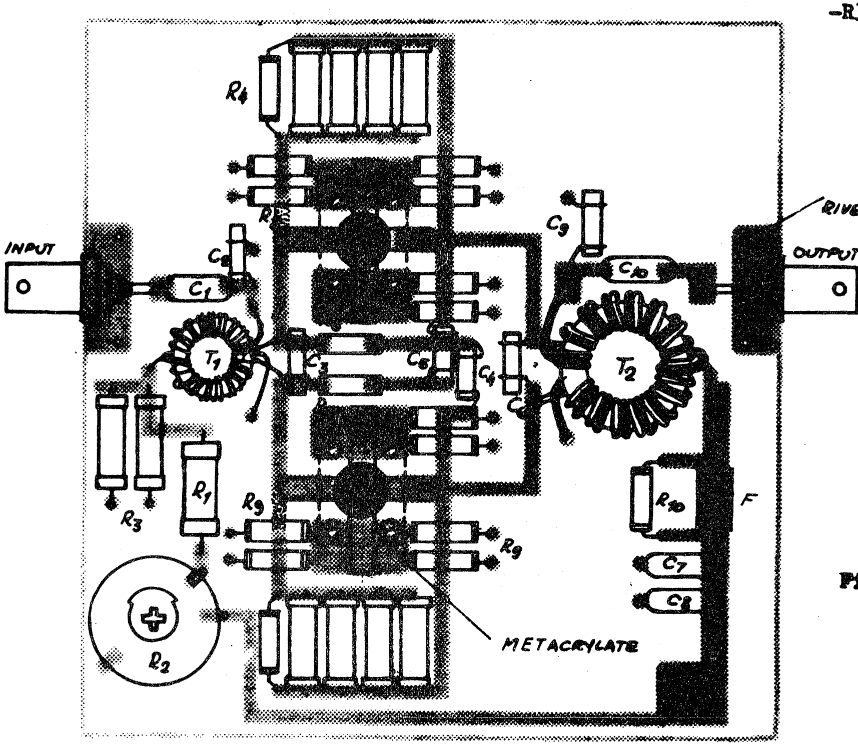


Fig. 27a

-UPPER SIDE-

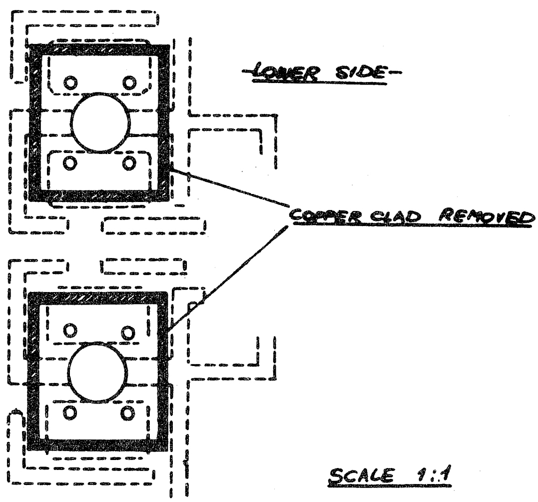


Fig. 27b

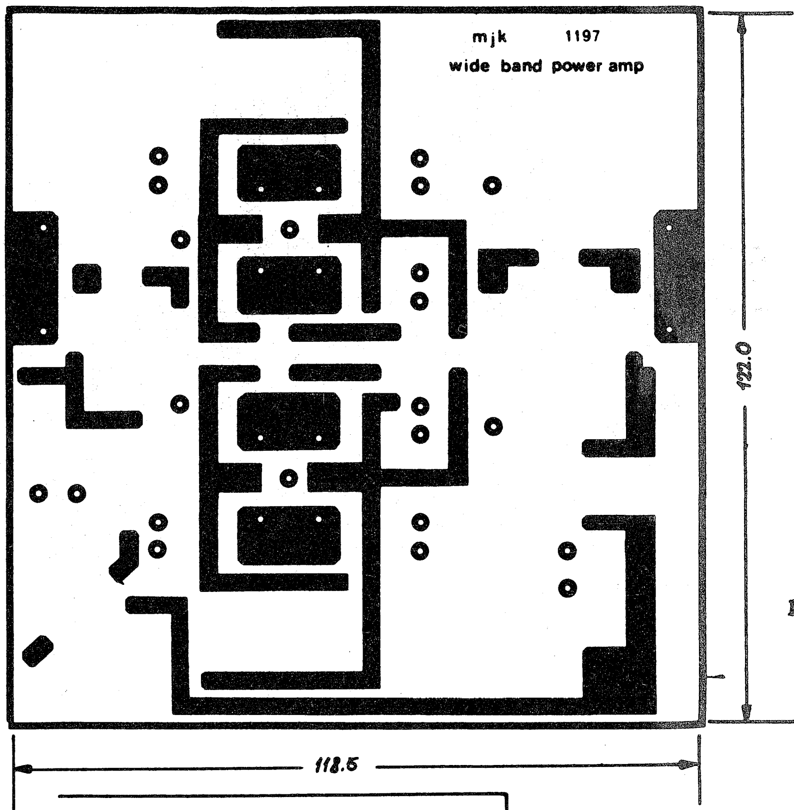
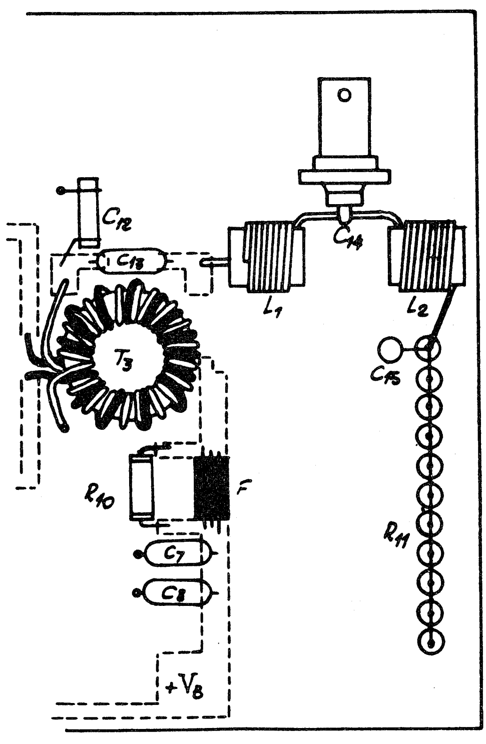


Fig. 27c



MODIFICATION TO
'TUBE DRIVER'

Fig. 27d

ELECTRONIC APPLICATION LABORATORY REPORT

GROUP : **C.A.B.**
Communication
EINDHOVEN

Report nr : ECO 7213
Date : 20.10.1972
Project nr : 6095
Pages : S1 + N0 + R10

AUTHOR : A.H. Hilbers

TITLE : Design of H.F. Wideband Power Transformers
Part II

SUMMARY

In part I of this report (ECO6907) the transmission line transformer has been discussed extensively. In this second part attention has been paid to the conventional transformer as a wideband matching element in e.g. S.S.B. transmitters.
The main problem in these transformers is stray-inductance.
Some H.F. compensation methods are discussed and a design example is given.

A.H. Hilbers

	Advies Octrooi d.d: 7-11-72	~~AV~~	GV		B		BL
	Opgave Mamo d.d: 15-12-72	~~AV~~	~~GV~~	~~EV~~	B		BL
	DATUM: 23 okt. 1972	MAMO:					

DESIGN OF H.F. WIDEBAND POWER TRANSFORMERS (Part II)

1. Introduction

Report EC06907 (Ref.1) and Application Information 530 (Ref.2) have been devoted entirely to the design of transmission line transformers. This type of transformer has undoubtedly the advantage of the largest possible bandwidth. However it has also some drawbacks:

a. the impedance transformation ratio is restricted to $(n : m)^2$ in which n and m are small integers.

b. they are difficult to construct when a number of windings must be combined.

Therefore it is certainly worthwhile to consider the possibility of applying a conventional transformer if the abovementioned problems arise.

2. L.F. limitations of conventional transformers.

These limitations are the same as for transmission line transformers, viz. parallel inductance and maximum flux density. They have been described in the abovementioned reports including compensation methods for the parallel inductance.

3. H.F. limitation of conventional transformers.

The most important property that limits the H.F. performance of a conventional transformer is the well-known stray-inductance. The equivalent circuit of the transformer for this frequency region is shown in fig. 1.

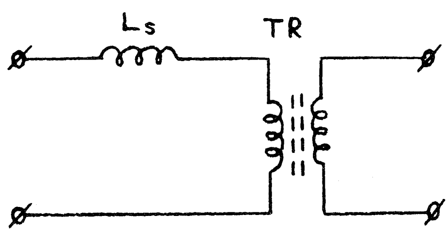


Fig.1

L_s is the stray-inductance and TR an ideal transformer. An easy way of measuring L_s is to measure the reactance between the primary terminals when the secondary is short-circuited. This can be done e.g. with the H.P. Vector Impedance meter. The most accurate result is obtained when the measurement is made at the highest frequency of operation and on the high-ohmic winding.

It is obvious that L_s must be kept as small as possible to avoid degradation of the H.F. performance of the transformer. For this end the following measures are recommended:

a. the windings must be as close to the core and to each other as possible.
b. each winding must be divided equally around the whole periphery of the core.
c. each winding must cover the core as much as possible.

Some practical steps that can be taken are:

1. the use of copper foil for the low-ohmic winding; this can be in direct contact with the core as the resistivity of the ferrite is very high. For better covering of the core two windings can be connected in parallel in such a way that one winding is wound in between the other one; the isolation material required in this case must be very thin.
2. for the high-ohmic winding enamelled copper wire can be used. An appreciable reduction of L_s can be obtained by parallel connection of two or more wires.

4. H.F. compensation of conventional transformers

In Ref. 3 some forms of compensation have been described. They are compared in the table below.

Number of compensation elements	0	1	2
Maximum X/R	0.18	0.44	1.09

Table I

The table is based on a maximum input V.S.W.R. of 1.2. The quantity X/R is the reactance of the stray-inductance (referred to the primary) divided by the nominal input resistance of the transformer. Compensation with one element is possible either by connecting a capacitor in parallel with the primary or by connecting a capacitor in parallel with the secondary winding.

Compensation with two elements is carried out by connecting capacitors in parallel with both primary and secondary windings. This method has already been applied several times. See reports: EC07113, EC07201 and EC07203. A short description will be given with reference to fig. 2.

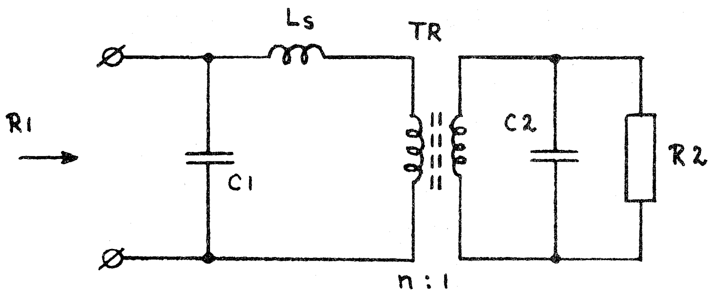


Fig. 2

TR is an ideal transformer with a voltage transformation ratio of n : 1 ($n > 1$) and L_s is the stray-inductance (referred to the primary). R_1 is the nominal input resistance:

$$R_1 = n^2 R_2$$

First we determine the normalized stray-inductance:

$L_{sn} = \dfrac{\omega_{max} L_s}{R_1}$, in which ω_{max} must be equal to or higher than 2π times the maximum frequency to be handled.

With the aid of fig. 3 we find the maximum input V.S.W.R. (S) and the normalized correction capacitance (C_{1n}). Then C1 and C2 can be calculated:

$$C1 = \frac{C1n}{\omega_{max} R_1} \quad \text{and} \quad C2 = n^2 C_1$$

A practical problem that can arise is that the capacitor across the low-ohmic winding must have such a high value that it approaches series resonance with its own lead inductance. In that case a sufficiently large number of smaller capacitors must be connected in parallel.

The practical limit to which the above described compensation system can be used is appr.:

$$L_{sn} = 1$$

If the stray-inductance is larger a more complicated compensation system is required. The above described system is based on a 3-element Chebyshev low-pass filter, but this can be extended to a 5-element network, which means that for the compensation 4 elements are required. This situation is depicted fig. 4.

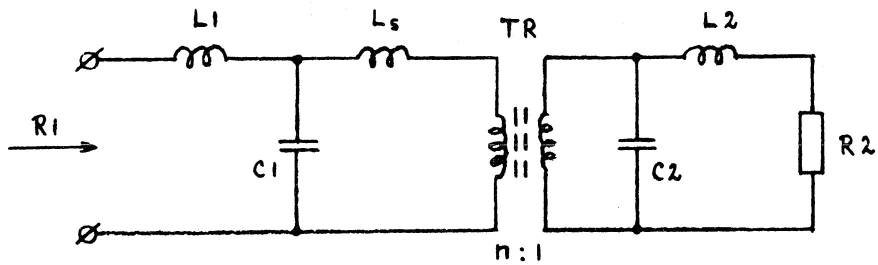


Fig. 4

Two inductors (L1 and L2) have been added. The normalized elements versus the maximum input V.S.W.R. (S) have been given in fig. 5*). From this graph it can be seen that L_{sn} may be as high as 1.77 for S_{max} = 1.2. A good practical limit is: $L_{sn} = 1.6$

*This graph and that of fig. 3 have been made with the aid of a computer program which is based on the design formulae for Chebyshev low-pass filters as can be found e.g. in Ref. 4.

The procedure for calculating the compensation elements is very similar to the previous case. We start again with determination of the normalized stray-inductance:

$$L_{sn} = \frac{\omega_{max}\ Ls}{R_1}$$

With the aid of fig. 5 we find the maximum value of S and the normalized values of the correction elements C_{1n} and L_{1n}. Now C1 and L1 can be calculated:

$$C_1 = \frac{C_{1n}}{\omega_{max} R_1} \quad \text{and} \quad L_1 = \frac{L_{1n} R_1}{\omega_{max}}$$

At the secondary side the correction components become:

$$C_2 = n^2 C_1 \quad \text{and} \quad L_2 = \frac{L_1}{n^2}$$

In general L_2 is so small that the tracks on the p.c. board can perform this function.

5. A practical example

For a low voltage S.S.B. amplifier a transformer was required with the following specification:

Frequency range	:	1.6 - 28 MHz
Power handling	:	52 W
Load impedance	:	100 ohms
Input impedance	:	4.6 ohms

The most suitable Ferroxcube material for this frequency range is 4C6. From the power handling point of view a toroid core will be chosen with the dimensions: 23 x 14 x 7 mm^3, catalog nr. 4322 020 91070.
The parallel reactance at 1.6 MHz measured at the secondary side must be +j400 ohms (see ref. 1 and 2), corresponding with an inductance of 40 μH.
The required number of turns is then:

$$n_{sec} = \sqrt{\frac{L.l}{\mu_o\ \mu_r A}}$$

in which $\frac{A}{l}$ = 0.5525 mm for this core and μ_r = 120 for 4C6 material, so:

$$n_{sec} = 21.9$$

A logical choice would be $n_{sec} = 22$. The voltage transformation ratio is $\sqrt{\frac{100}{4.6}} = 4.66$, by which the primary number of turns must be: 22/4.66=4.7. We choose: $n_{pr} = 5$.

The required transformation ratio can be approached better if the secondary number of turns is modified from 22 to 23. Now the secondary parallel inductance becomes 44 µH and the impedance transformation ratio $\left(\frac{23}{5}\right)^2 = 21.1$ by which the input impedance will be 4.74 ohm being appr. 3% higher than the required value.

L.F. compensation of the parallel inductance can be carried out by means of a single capacitor in series with the secondary (see ref. 1 and 2):

$$C_L = \frac{L_{psec}}{R^2_{sec}} = \frac{44 \times 10^{-6}}{100^2} = 4400 \text{ pF}$$

A standard value of 4700 pF has been chosen.

The maximum voltage across the secondary winding is:

$$V_{sec} = \sqrt{2R_{sec}P} = \sqrt{2x100x52} = 102V$$

Now the maximum flux density can be calculated:

$$B_{max} = \frac{V_{sec}}{\omega_{min} A \, n_{sec}}$$

in which A = 31.5 mm² for this core, so:

$$B_{max} = \frac{102}{2\pi \ x1,6x10^6x31,5x10^{-6}x23} = 140 \text{ gauss at } 1.6 \text{ MHz.}$$

This gives a core loss of appr. 1% or 0.5W.

To keep the stray-inductance low the transformer has been wound as follows:

- The primary consists of the parallel connection of two windings each having 5 turns of 4 mm wide copper foil. Each winding has been equally divided around the periphery of the core; one winding was wound in between the other one. Between these two windings a thin layer of isolation material was used.

- The secondary consists of the parallel connection of two windings each having 23 turns of 0.45 mm enamelled copper wire. The method of winding was the same as for the primary.

The stray inductance measured at the secondary side was 0.67 μH. To make the correction less critical we choose a maximum frequency of 35 MHz instead of the specified 28 MHz. The normalized stray-inductance becomes then:

$$L_{sn} = \frac{2\pi \times 35 \times 10^6 \times 0.67 \times 10^{-6}}{100} = 1.47$$

From fig. 5 we find:

$S_{max} = 1.064$

$C_{1n} = 1.24$

$L_{1n} = 0.66$

In this case the index 1 applies to the secondary and index 2 to the primary. The compensation elements become:

$$C_1 = \frac{1.24}{2\pi \times 35 \times 10^6 \times 100} = 56.4 \text{ pF}$$

$$L_1 = \frac{0.66 \times 100}{2\pi \times 35 \times 10^6} = 0.3 \ \mu\text{H}$$

C2 = 21.1 x 56.4 = 1190 pF
L2 = 0.3/21.1 = 0.0142 μH

For C1 a standard value of 56 pF was chosen.
C2 consisted of the parallel connection of 330pF and 3x270 pF being 1140 pF in total; the slightly lower value was chosen because of the series inductance of the capacitors. L2 was formed by the tracks on the p.c. board.

The first measurements on the compensated transformer showed a too high V.S.W.R. at 28 MHz together with a capacitive behaviour of the impedance from which we got the impression that it was over-compensated. This appeared to be due to the capacitance between the primary and the secondary winding. Therefore we reduced the values of the compensation elements by 10 to 20 %. The new values are:

C1 = 47 pF
L1 = 0.27 μH
C2 = 980 pF (parallel connection of 2 x 270 pF and 2 x 220 pF)
L2 ≈ 2 x 6 nH (tracks on p.c. board)

The complete situation is depicted in fig. 6. The load impedance Z_L consisted of the parallel connection of 2 resistors of 15 ohm, 1 resistor of 12 ohm and a capacitor of 120 pF. The latter was added to compensate the series inductance of the resistors.

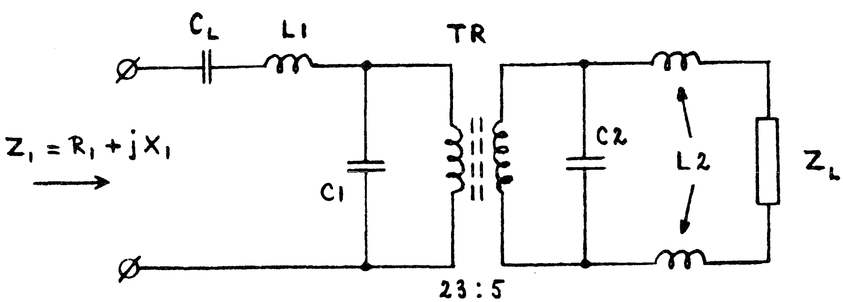


Fig. 6

The results of the measurements have been summarized in table II

f (MHz)	R1 (ohm)	X1 (ohm)	V.S.W.R. (-)
1.6	98	- 1.87	1.03
3.5	100.5	- 2.1	1.02
7.0	98.3	- 5.84	1.06
14	89.9	- 4.41	1.12
20	87.4	+ 1.24	1.14
28	100	+ 7.18	1.07

Table II

It can be seen from the table that the maximum input V.S.W.R. has been reduced to 1.14. Without H.F. compensation this V.S.W.R. would have been appr. 3 at 28 MHz.

A.H. Hilbers

References:

1. A.H. Hilbers; "On the Design of H.F. Wideband Power Transformers", C.A.B. report nr. EC06907, March 14, 1969.

2. A.H. Hilbers; "Design of H.F. Wideband Power Transformers", Application Information nr. 530, June 17, 1970.

3. H. Nielinger; "Optimale Dimensionierung von Breitbandanpassungsnetzwerken", N.T.Z. 1968, Heft 2, pp. 88–91.

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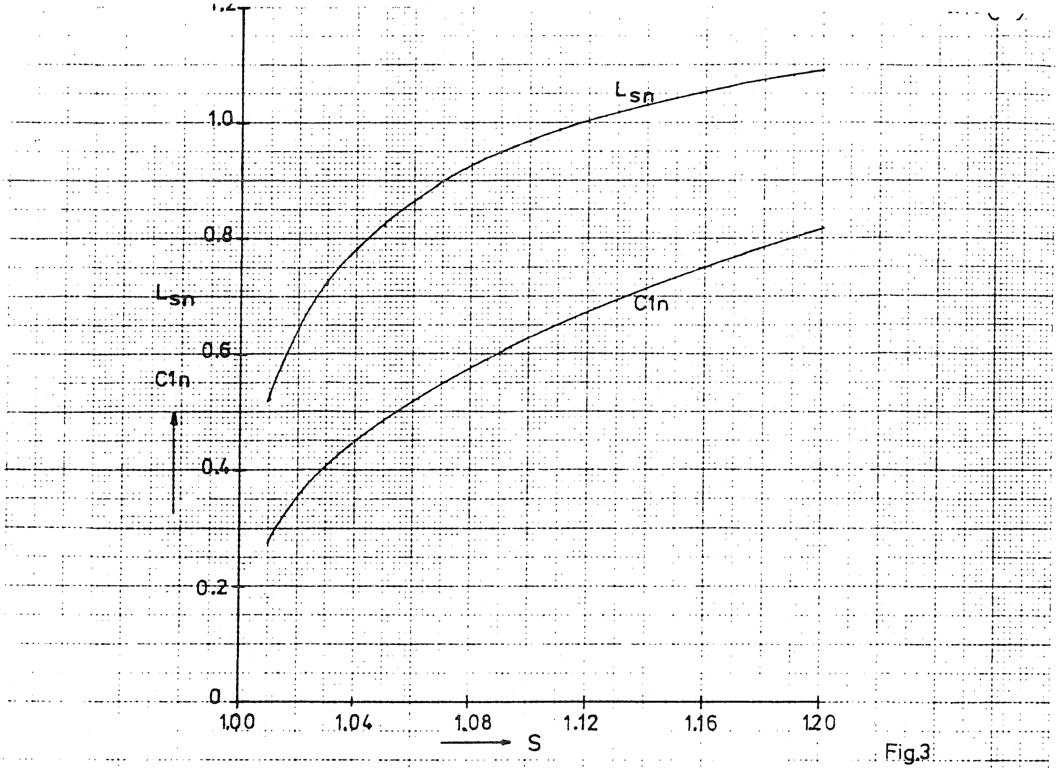
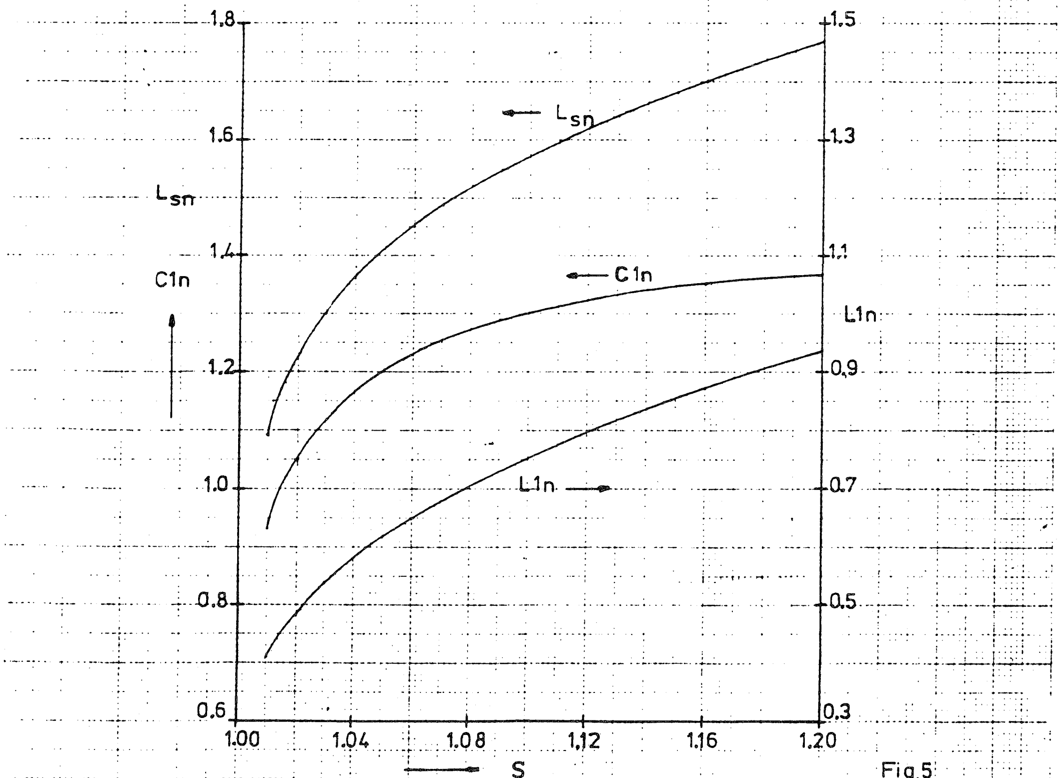


Lsn
C1n
Lsn
C1n
0
0.2
0.4
0.6
0.8
1.0
1.00
1.04
1.08
1.12
1.16
1.20
S

Fig.3



Lsn
C1n
Lsn
C 1n
L1n
L1n
0.6
0.8
1.0
1.2
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1.6
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0.5
0.7
0.9
1.1
1.3
1.5
1.00
1.04
1.08
1.12
1.16
1.20
S

Fig.5

ELECTRONIC APPLICATION LABORATORY REPORT

GROUP : **C.A.B.**
TELECOMMUNICATIONS
EINDHOVEN

AUTHOR: M.J. Köppen

Report nr: ECO 7308
Date : 19-10-1973
Project nr: 6095
Pages : P1 + A4 + R47

TITLE : A SINGLE STAGE WIDEBAND (1.6 - 28MHz) LINEAR POWER AMPLIFIER FOR 300 WATTS PEP USING 2x BLX15.

PURPOSE and SCOPE

- This report is primarily intended for the manufacturers of S.S.B. transmitters in the H.F. range (1.6 - 28MHz).
- The objective of this project was the design of a wideband power amplifier module of which a combination of four is capable to supply 1KW output power.
- This module contains 2 transistors type BLX15 connected in push-pull and adjusted in class - AB.
  At a supply voltage of 50V an output power can be delivered of 30 - 300W P.E.P. at an intermodulation distortion of -30dB or better.
  In the frequency range of 1.6 - 28MHz the power gain is equal to 16.8 ± 0.5dB; the input V.S.W.R. has a maximum value of 1.17.
- The only competitive device in this field is the C.T.C. transistor type S150 - 28 which delivers 150W output power at a supply voltage of 28V. This is not so attractive from the application point of view because of the very low impedance levels.

G. Fink
A.H. Hilbers

Advies Octrooi d.d.: 30 okt. 1973	~~AV~~	GV		B		BL
Opgave Mamo d.d.: 2 nov 1973	~~AV~~	~~GV~~	~~BI~~	B		BL
DATUM: 19 okt. 1973	**MAMO**:					

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ABSTRACT

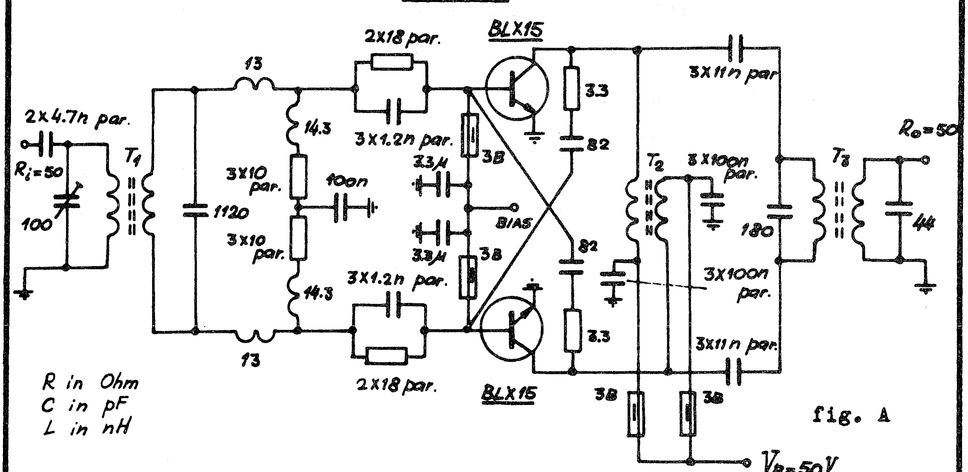

fig. A

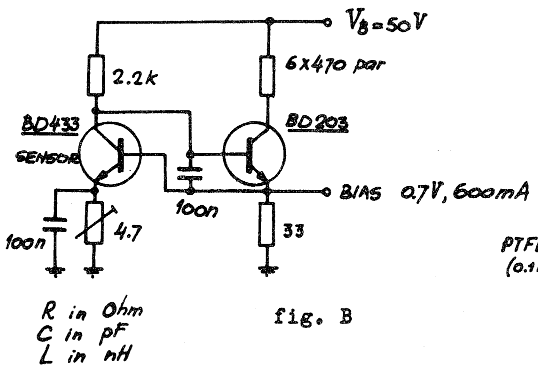

fig. B

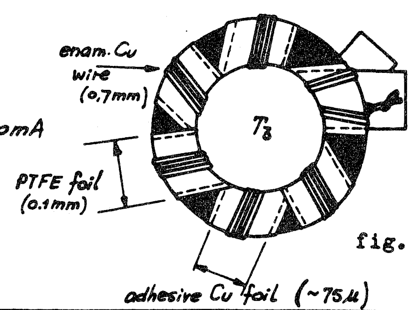

fig. C

	Advies Octrooi d.d.:	AV	GV		B		BL
	Opgave Mamo d.d.:	AV	GV	EI	B		BL
	DATUM: 19 okt 1973	MAMO:					

Figure A shows the schematic diagram of the wide band power amplifier. The linear transistor BLX15 (development type nr. 402 BLY) is used in push-pull configuration.

The design has been started with the calculation of the intrinsic parameters of the transistor from the mask drawings, material specification, etc.

The large signal behaviour then could be calculated with a computer program as described in earlier C.A.B. reports.
These calculations show large variations of the input impedance and gain when operating over more than four octaves.

The problem has been solved by inserting a six elements network per transistor, comprising the components between the input transformer T_1 and both BLX15s and by applying cross-neutrodynisation. It was possible to calculate the network elements with the aid of an available computer program.

A good compromise has been found in the choice of a network input impedance being equal to 2 x 2,78 = 5,56 Ohms, a value that easily can be transformed to the 50 Ohms source impedance by means of T_1, being an asymmetrical to symmetrical 9:1 impedance transformer.

The required load impedance of a pair of BLX15s amounts to appr. 12,5 Ohms, so a transformer ratio 1:4 has been chosen.

Both transformers have been constructed as conventional types wound on 4C6 toroids. The input transformer uses a toroid with dimensions Dxdxl= $14x9x5mm^3$, whilst calculations proved that the application of our biggest toroid of 4C6 material, with dimensions Dxdxl= $36x23x15mm^3$ can be justified for T3, what has been confirmed by practical tests.

Fig. C gives an example of the practical realisation of the output transformer. The construction combines the advantages of high mutual coupling, small parasitic capacitance and low ohmic losses.

To reduce the influence of the spreading inductance, particularly at the high end, and shunt inductance at the lower end, compensation has been applied according to the methods being already described in preceding C.A.B. reports concerning this subject.

The output circuit contains a centre tapped choke coil T2, that consists of an FXC rod of 4B1 material with windings of twisted enamelled wire.

../A3

In the cross neutralisation chain parallel connected resistors, with a resulting value of 3,3 Ohms, have been inserted to prevent oscillations in case of insufficient output matching.

The amplifier operates from a 50 Volts supply voltage, what has consequences for the applied bias network (Fig. B).
As emitter follower the NPN silicon epitaxial base transistor BD203 can do the job, whilst the temperature sensor is a BD433.

The complete amplifier is mounted on a p.c. board, being screwed against a water cooled heatsink.

The transistors must form a matched pair.

In total three amplifiers have been built. The typical results from single tone measurements of one amplifier for an output of 300 Watts are shown in Fig. D.

The intermodulation is measured with respect to one tone of a double tone test signal with a frequency difference of appr. 1kHz at output powers of 30, 100, 200 and 300 Watts PEP.

Figs. E and F show the d3 versus frequency. The value of the d5 is small enough with respect to d3 and has not been mentioned in the curves.

../A4

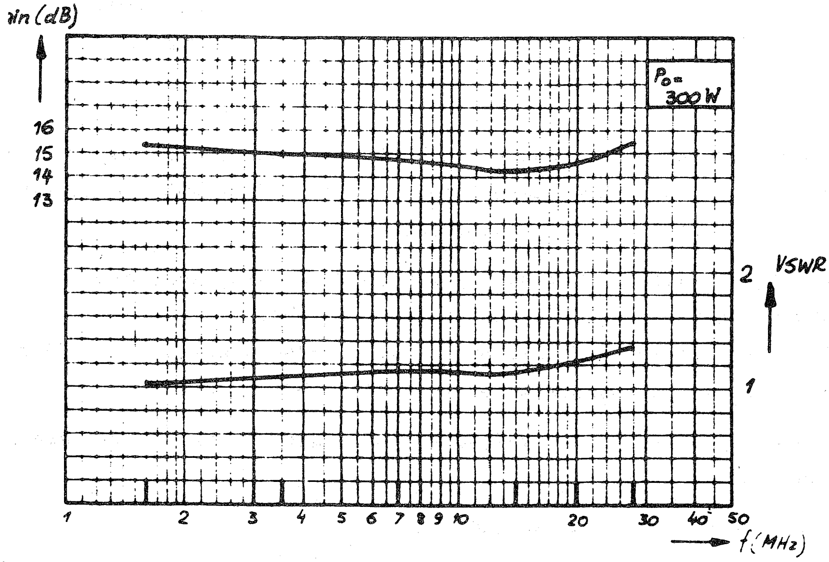

fig. D

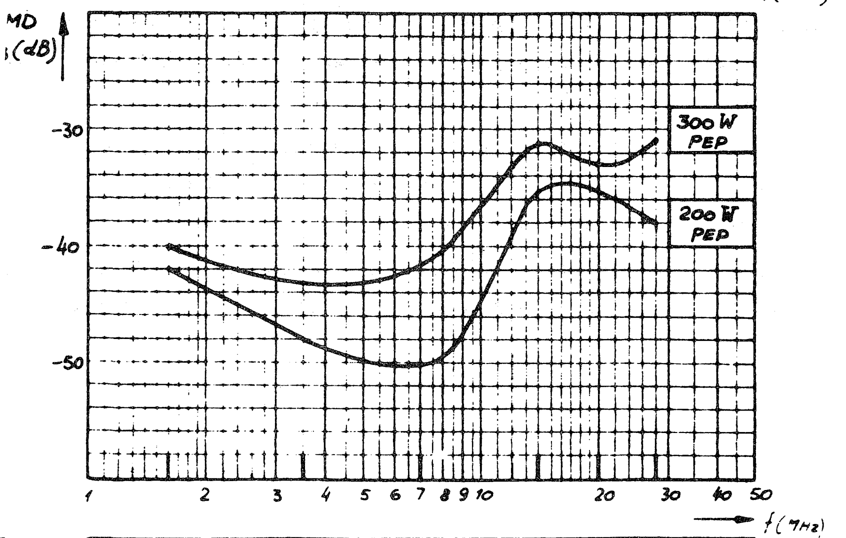

fig. E

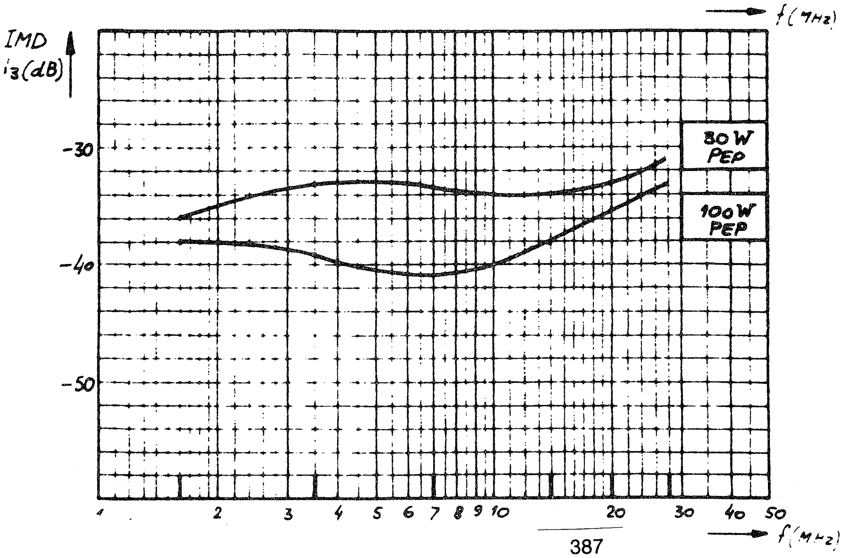

fig. F

1. INTRODUCTION

This report describes a wide-band linear power amplifier for the 1,6 - 28MHz frequency range. The design is based on the BLX15, a silicon epitaxial NPN transistor specially designed for industrial and military SSB and RTTY applications in this band.

The wide-band amplifier uses two BLX15s in a push-pull configuration with an output power of 300 Watts PEP.

An intermodulation distortion distance of at least -30dB could be realised.

The input impedance of the amplifier has a low VSWR because it will form the load of the preceding driver stage.

../2

2. CIRCUIT DESCRIPTION

The applied transistor is the BLX15 (development number 402BLY) intended to work from a supply voltage of 50 Volts.

The design has been started with the calculation of the intrinsic parameters of this type from the mask drawings, material specification etc.

The large signal behaviour then could be calculated with the computer program as described in C.A.B. reports ECO 7112 (Ref. 1) and ECO 7118 (Ref. 2) or in Electronic Applications (Ref. 3 and Ref. 4).

Calculations were done with a program written in CPS (Conversational Programming System) on an IBM 360/75.

With the mentioned program it is possible to calculate $R_i \pm jX_i$, R_{load} and C_{load}, the input power P_i and the power gain as a function of frequency, output power Po and supply voltage Vcc.
At the same time it is possible to choose the class of operation (A, B or C), the phase angle between collector voltage and current (see Ref. 1 or 3) and the amount of neutrodynisation.

In the first instance we started with the development type 402BLY. The table below mentions the input data:

f_T= 250MHz	C_{be} = 1970pF	P_{load}= 150 Watts
h_{fe}= 30	C_{cbi}= 38,9pF	V_c = 46,8 Volts
R_b= 0,18 Ohm	C_{cbo}= 86,9pF	V_{bo}= 0,7 Volt
R_e= 0,14 Ohm	C_{ce} = 69,7pF	R_{bo}= 0,2 Ohm
R_c= 0,37 Ohm	C_s= 3,5pF	R_{be}= 10^{12} Ohms
L_b= 1,7nH		R_{bc}= 10^{12} Ohms
L_e= 1nH		C_n= 82pF
L_c= 1,5nH		

The calculations were done on an equivalent circuit as shown in Fig. 1.

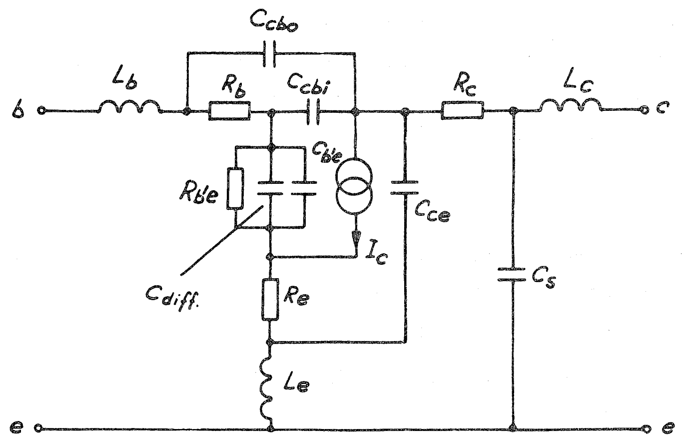

Fig. 1

The next table shows the results of the calculations on the 402BLY. They were done at some characteristic frequency points in the band 1,6 - 28MHz.

f (MHz)	R_L (Ohm)	C_L (pF)	R_i (Ohm)	X_i (Ohm)	Pi (Watt)	G (dB)
1.6	6.26	-273.1	7.47	-2.77	0.235	28.05
2.5	6.26	-273.1	6.28	-3.53	0.241	27.94
3.5	6.26	-273.1	5.06	-3.81	0.251	27.75
5	6.26	-273.2	3.71	-3.64	0.273	27.39
7	6.26	-273.2	2.65	-3.12	0.315	26.78
10	6.25	-273.4	1.88	-2.40	0.403	25.70
14	6.23	-273.7	1.44	-1.73	0.570	24.20
20	6.20	-274.3	1.19	-1.11	0.927	22.10
24	6.18	-274.8	1.12	-0.83	1.236	20.85
28	6.15	-275.4	1.07	-0.61	1.605	19.72

As can be seen from these results the input impedance and the gain of a transistor operating over more than four octaves show large variations. The input power needed for 150 Watts output increases from 0,235 to 1,605 Watts what results in a gain variation of 28,05 - 19,72 = 8,23dB.

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The complex input impedance $R_i \pm jX_i$ shows a large variation too, whilst the R_{load} and C_{load} may be supposed to be constant, what means that the transistor has to be loaded with 6,25 Ohms in parallel with an "inductance" of -274pF.

The load impedance was chosen equal to 6,25 Ohms to keep the possibility of applying a transmission line type transformer, because the property of this type is that the impedance transformation equals $1:n^2$, in which n is a small integer. Within certain limits this value can be chosen by the required output power and the amplitude of the collector voltage V_c. The relation is given by the approximation: $R_{load} \approx V_c^2/2.P_o$

The target is 300 Watts output in linear operation with an intermodulation distance of at least -30dB. From two pieces BLX15 operating in class AB it can be expected that they can do the job.

Adding power can be arranged by means of a parallel or a push-pull configuration. We have chosen for the push-pull system with cross neutralisation. The reason why can be found in C.A.B.report ECO 7114 (Ref. 5) describing a similar set-up with two BLX14s for 80 - 100 Watts PEP output.

The output power generated by the transistor pair has to be delivered to a 50 Ohms load. Because the values of R_{load} and C_{load} are almost constant a wide band output transformer with ferrite core is possible.

Taking in view the high power and the fact that this transformer may not contribute too much in the overall distortion the construction needs special attention.

Two versions are possible viz. the transmission line type and the conventional type. The reason why the latter wide-band type has been chosen will be explained further on.

The input power presented to the circuit has to be rather constant versus frequency, whilst the input VSWR may not become too high, preventing mismatch of the preceding driver stage.

It is desired to keep the input VSWR below 2:1 and the gain variation within $\pm$ 1,0dB.

For this reason a wide-band input transformer can only be applied when the load is rather constant versus frequency and has a suited value.

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The variations in input impedance and gain have been decreased in two ways:

a) Applying cross neutralising from the base of one transistor to the collector of the other and vice versa. A value of 82pF is suitable.

b) Applying a correction network at the input. The purpose and motivation of such a network has been described extensively in report ECO 7114 (Ref. 5).

Fig. 2 shows a simplified correction network of one transistor input.

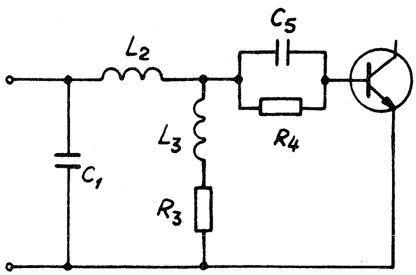

Fig. 2

Because this network is still rather theoretical it was decided to incorporate all parasitic inductances and loss resistances and to take if possible available standard values for the resistors and capacitors.

The new extended circuit takes the form of Fig. 3

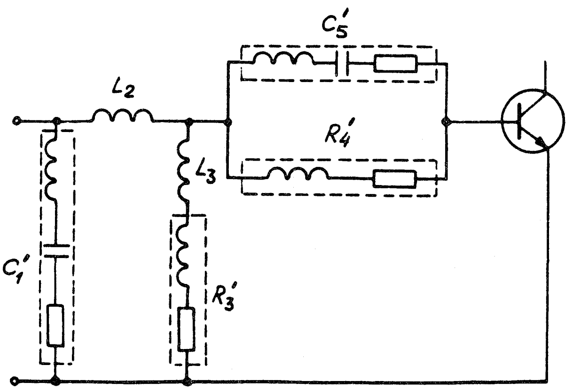

Fig. 3

../6

The parasitic elements of the components are known from experience and have been checked in the past.
The approximate values for these elements of a polystyrene capacitor are 8nH for the inductance and 0,1 Ohm for the loss resistance. The leads of a capacitor were appr. 2x3mm. With the same lead length the resistors have also a parasitic inductance of 8nH. The losses in L_2 and L_3 are neglected.

The calculation of the network components of Fig. 2 for approximately equal gain and required input impedance at the two outer frequencies 1,6 and 28MHz can be done with the calculation method of report ECO 7114 (Ref. 5).
To know more details about the intermediate frequencies a program has been used made for the IBM 360/75 computer.

It calculates, among other things, the overall input impedance and gain as functions of frequency. A number of 10 frequencies was chosen mostly at half octave points.
The transistor data at these points must be introduced in the program.
The results of the first approximation have been given in the following table of component values:

C_1 = 1113.9pF , L_2 = 13,3nH , R_3 = 3,32 Ohms
C_5 = 3817.5pF , L_3 = 20,3nH , R_4 = 9,11 Ohms

When these calculations were done it turned out that the maximum input VSWR occurred at 20MHz being 1.18:1.
The power gain is calculated with the input mismatch taken into account:
$G = 10 \log (P_o/P_{if})$ in which P_o is the output power and P_{if} the incident drive power.

The gain was maximum at 1,6MHz viz. 16,88dB and had a minimum of 16,48dB at 7MHz so a variation of $\pm$ 0,21dB. These values are within the requirements which were for the VSWR a maximum of 2:1 and for the gain $\pm$ 1dB.

Now all parasitic elements of Fig. 3 have been introduced in the calculation, the values of the coils were rounded off and standard values were taken for resistors and capacitors. Simultaneously the value of L_3 has been reduced to 17nH to decrease the gain variation versus frequency.

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The new values are listed below:

$C_1' = 2 \times 560pF$ parallel + 4nH + 0,05 Ohm

$L_2 = 13nH$

$L_3 = 14,3nH$

$R_3' = 3 \times 10$ Ohms parallel + 2.7nH

$R_4' = 2 \times 18$ Ohms parallel + 4nH

$C_5' = 3 \times 1200pF$ parallel + 2,7nH + 0.03 Ohm

Calculations executed with these values gave a slightly lower VSWR of 1.14:1. The gain variation increased from ± 0,21dB to ± 0,4dB (min. 16,24dB, max. 17,02dB).

The next table shows in short form the output data of the computer.

f (MHz)	R_{is} (Ohm)	X_{is} (Ohm)	VSWR (-)	$P_{i_{abs.}}$ (Watt)	Gain (dB)
1,6	2,81	0,01	1,01	3,063	16,90
2,5	2,81	0,01	1,01	3,177	16,74
3,5	2,82	0,01	1,02	3,307	16,57
5	2,85	0,01	1,02	3,461	16,37
7	2,89	-0,01	1,04	3,562	16,24
10	2,93	-0,09	1,06	3,546	16,26
14	2,91	-0,23	1,10	3,376	16,47
20	2,67	-0,29	1,12	3,084	16,86
24	2,56	-0,06	1,09	2,973	17,02
28	2,75	0,36	1,14	2,975	17,01

2.1 CHOICE OF TRANSFORMER TYPE

The performance of the broadband amplifier is determined to a large extent by the quality of the transformers used in the circuit. In the frequency range concerned careful design is essential if low losses and accurate impedance transformation are to be maintained over the whole band.

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Broadband transformers can be constructed in the conventional way, with a magnetic core and separate primary and secondary windings, or as a transmission line type in which the primary and secondary windings are combined to form one or more transmission lines.

C.A.B.report ECO 6907 (Ref. 6) and Application Information 530 (Ref. 7) have been devoted entirely to the design of the transmission line transformer. The latter, having undoubtedly the advantage of the largest possible bandwidth, has also some drawbacks:

a) the impedance transformation ratio is restricted to $1/n^2$ compared with n^2/m^2 for the conventional type (n and m are small integers)
b) they are difficult to construct when a number of windings must be combined.

However, they have very low insertion losses, whilst the winding capacitances and stray inductance, which set the H.F.limits of convention transformers, are absorbed in the characteristics of the lines. The L.F. limits of both versions are the same.

NiZn ferrite (a.o. grade 4C6) is the most suitable core material and the toroid the most useful core shape.
Mainly the points a and b led to the choice of the conventional type.

2.2 THE INPUT TRANSFORMER

The input transformer has to transform the 50 Ohms asymmetrical source impedance to the 2 x 2,78 = 5,56 Ohms symmetrical network input impedance So the transformer is a 9:1 impedance transformer.

The reactance of the shunting inductance at the lowest frequency 1.6MHz has been chosen to be at least 4 times the primary resistance (50 Ohms), what roughly corresponds with an input VSWR of $1+ R_i/X_L = 1,25$ a value being improved by compensation later on.

So the inductance is 20 μH or +j200 Ohms at 1,6MHz.

The transformer is wound on a ferrite toroid of 4C6 material.
These toroids are delivered in standard dimensions and it must first be investigated which size is needed.

This can be done with the aid of the formulae:

$L = \mu_o . \mu_r . n^2 . A/l$ and $B_{max} = V_{max}/\omega . A . n$

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A is the average ferrite cross section in m^2 and l the average length of the lines of force in m. Both A and l are unknown.

Finding the number of turns n from the first equation and introducing it in the second one gives after rearranging

$$A.l = (V_{max}/\omega.B_{max})^2 . \mu_o . \mu_r / L$$

in which: $\mu_o = 4.\pi .10^{-7}$

$\mu_r = 120 \pm 20\%$ (4C6)

$L = 20\,\mu H$

$\omega = 2\pi \times 1.6 \times 10^6$ rad/sec.

V_{max} can be calculated from the power that has to be transferred. The input power in the push-pull configuration amounts to approximately 7,5 Watts for 300 Watts output and 16dB of gain.

In that case: $V_{max} = (2.P_i.R_g)^{\frac{1}{2}} = (2x7.5x50)^{\frac{1}{2}} = 27{,}4$ Volts

B_{max} depends on the parallel loss resistance that is acceptable what can be calculated from the Figs. 4 and 5.

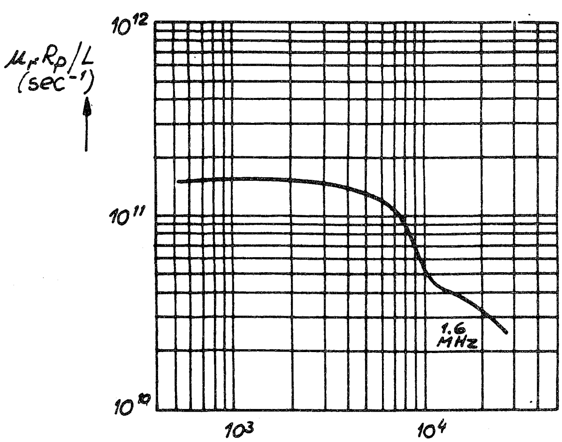

Fig. 4 ⟶ $B_{max} \cdot f$ (T.Hz)

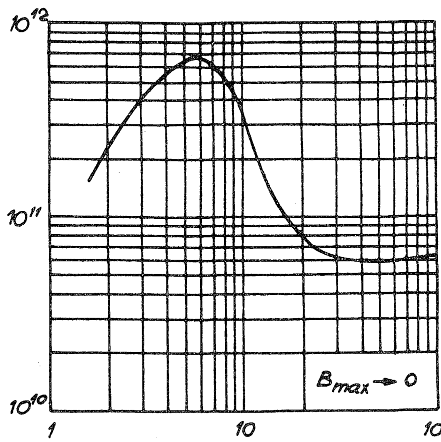

Fig. 5 ⟶ f (MHz)

Fig. 4 shows μ_r Rp/L versus $B_{max}.f$ for 1,6MHz and Fig. 5 μ_r Rp/L versus f for $B_{max} \rightarrow 0$.

If a power loss of 1% is acceptable then Rp = 100 x 50 = 5000 Ohms. Hence $\mu_r.Rp/L = 120x5x10^3/20x10^{-6} = 3x10^{10}$ sec $^{-1}$. As at high frequencies Rp depends hardly on B_{max} we may conclude from Figs. 4 and 5 that above calculated value only can occur at 1,6MHz.

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The value of $B_{max.}f$ corresponding with above mentioned U_r, Rp/L amounts to $2x10^4$T.Hz according to Fig. 4, so the resulting value for $B_{max.}$= $2x10^4/1,6x10^6$ = $1,25x10^{-2}$T at 1,6MHz.

The required product A.l can now be determined:

A.l.= $(27,4/2\pi x1,6x10^6x1,25x10^{-2})^2x4\pi x10^{-7}x120/20x10^{-6}$= $0,36x10^{-6}m^3$

In the table below the A.l products of some available toroids have been given, calculated according to:

A.l = $\left\{h\ (D-d)/2\right\}\left\{\pi (D+d)/2\right\}$

catalog number	Dimension (mm) Dxdxh	$A(m^2)$	$l(m)$	$A.l(m^3)$
4322 020 91010	9x6x3	$4,51x10^{-6}$	$2,33x10^{-2}$	$0,105x10^{-}$
4322 020 91020	14x9x5	$12,54x10^{-6}$	$3,55x10^{-2}$	$0,445x10^{-6}$

As the table shows the smallest toroid that meets our needs is a type with the dimensions 14x9x5mm.
The number of turns for an inductance of 20 μH is calculated with the aid of the rearranged equation:

n= $(L.l/\mu_o.\mu_r.A)^{\frac{1}{2}}$= $(20x10^{-6}x3,55x10^{-2}/4\pi x10^{-7}x120x12,54x10^{-6})^{\frac{1}{2}}$= 19,35 turns.

The number of secundary turns depends on the impedance step to be made:
n_{prim}/n_{sec}= $(Rp/Rs)^{\frac{1}{2}}$= $(50/5,56)^{\frac{1}{2}}$= 3

It is obvious to take the ratio: 21:7 turns.

During experiments it appeared that a mid tap is superfluous, it could be arranged in an artificial way.

In first instance a transformer has been constructed having 21 turns primary and 7 turns secondary. The wire diameter was 0,6mm of enamelled copper wire.
To keep L_{Str} small the windings were as close together and to the core as possible. Also each winding was divided around the periphery of the core and wound in between the other one.

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With the aid of a vector impedance meter (HP4815A) the L_{Str} has been measured. It appeared that L_{Str} was too high to apply compensation according to Nielinger (Ref.8).
The factor $X_{Lmax}/R = 70{,}79/50 = 1{,}416$ at 28MHz.
This value must be reduced to below 1.1.

Then the idea arose to use adhesive copper foil (width 3mm; thickness 0,075mm) for the low-ohmic winding. This can be in direct contact with the core as the resistivity of the lacquered ferrite is very high.

The high-ohmic winding consisting of enamelled copper wire of 0,5mm has been wound over the secondary in such a way that 3 turns cover one turn of the secondary (see constructional details on page 28).
Between both windings a thin (width 5mm ; thickness 0,1mm) sheet of PTFE foil was required as isolation.

The L_{Str} measured at the two outer frequencies with short-circuited secondary is:

f(MHz)	X_{Str}(Ohm)
1,6	2,627
28	44,49

2.3 H.F. COMPENSATION OF THE INPUT TRANSFORMER

Fig. 6 shows the applied system seen from the primary.

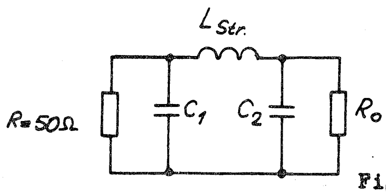

Fig. 6

Ro is the transformer output resistance and L_{Str} the stray inductance seen from the primary

According to Nielinger (Ref. 8) $C_1 = C_2$ will be added to form an L.P. section by which the maximum VSWR is reduced.
C_2 is transformed from the secondary.

../12

$X_{Lmax} = 2.\pi .f_{max}.L = 44{,}49$ Ohms

$f_{max} = 28MHz$

$X_{Lmax} /R = 44{,}49/50 = 0{,}89 \quad \therefore \quad VSWR_{max} = 1{,}07$

From the curve of (Ref. 8) it follows that:

$R/X_{c\ min} = 0{,}56 \quad X_{c\ min} = 50/0{,}56 = 89$ Ohms

$X_{c\ min} = 1/2\pi .f_{max}\ C_{prim}$

$C_{prim} = 1/2\pi\ x28x10^6x89 = 63{,}7pF$ whilst C_{sec} becomes $T^2.C_{prim} = 9x63{,}7pF = 574pF$.

In first instance the primary capacitance was chosen 62pF. Later on it has been replaced by a 100pF film dielectric trimmer to make a tuning procedure possible.

The partly compensated transformer has been checked with a vector impedance meter. First the parasitic properties of the instrument probe have been determined. The results have been put in a small digital computer, which output shows the performance versus f given in Xs, Rs, Rp, Xp and VSWR. This method will be applied in subsequent measurements.

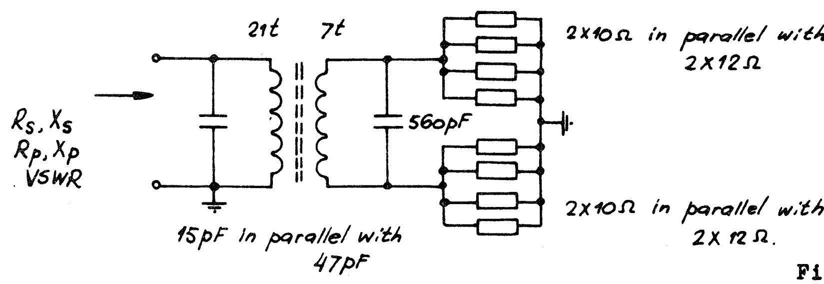

Fig. 7

f(MHz)	Rs(Ohm)	Xs(Ohm)	Rp(Ohm)	Xp(Ohm)	VSWR
1,6	49,59	9,58	51,44	266,34	1,21
3,5	50,30	3,65	50,56	697,25	1,08
7	50,96	0,24	50,96	10900	1,02
14	52,41	2,48	52,52	1108	1,07
20	51,50	5,47	52,08	490	1,12
24	49,72	6,66	50,61	377,6	1,14
28	48,60	7,04	49,61	342,6	1,16

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The compensation characterised by the VSWR at the high end of the band looks acceptable. At the lower end the VSWR (= 1.21) can be decreased by applying LF compensation.

2.4 L.F. COMPENSATION OF THE INPUT TRANSFORMER

Fig. 8 shows the essential elements for a simple high pass section at 1,6MHz.

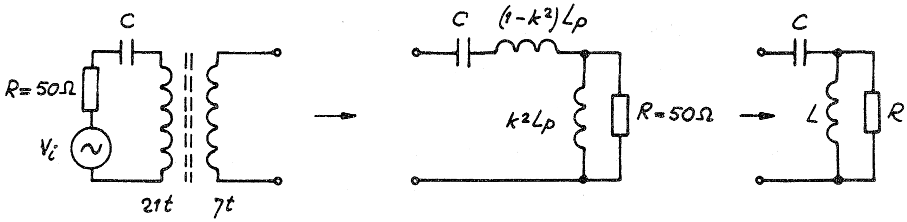

Fig. 8a Fig. 8b Fig. 8c

The primary inductance at f= 1,6MHz has been measured and calculated. The measured value is 26 μH. Because of spread the calculated value of 23.1 μH has been used.

According to ref. 8:

$R/X_{L\ min} = X_{c\ max}/R$

$X_{L\ min} = 232$ Ohms

$R/X_{L\ min} = 50/232 = 0,216 \quad \therefore \quad$ VSWR = 1,045

$X_{C\ max} = 0,216 \times 50 = 10,8$ Ohms

$C = 1/2\pi \cdot f_{min} \cdot X_{C\ max} = 1/2\pi \times 1,6 \times 10^6 \times 10,8 = 9,20$nF

A parallel connection of 2x 4,7nF has been chosen.

2.5 THE COMPENSATED INPUT TRANSFORMER

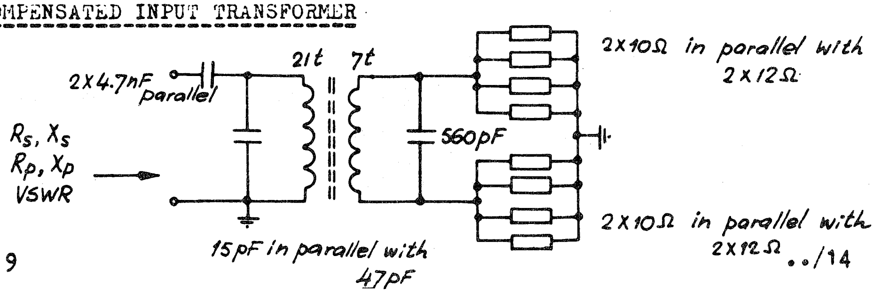

Fig. 9

../14

The table below shows the measured results:

f(MHz)	Rs(Ohm)	Xs(Ohm)	Rp(Ohm)	Xp(Ohm)	VSWR
1,6	49,52	0,72	49,53	3415	1,02
3,5	50,45	0,54	50,45	4748	1,01
7	50,95	0,69	50,96	3776	1,02
14	51,83	3,36	52,05	803	1,08
20	51,50	5,47	52,08	490	1,12
24	50,36	5,86	51,04	438	1,12
28	49,74	6,35	50,55	396	1,14

It shows a clear improvement of the VSWR at lower frequencies.

2.6 THE OUTPUT TRANSFORMER

The output transformer has the same lay-out as the input transformer but the calculation will be done in a different way because we approach the limit of what can be done with our biggest 4C6 toroid ($36x23x15mm^3$).

We start with the calculation of the required number of secondary turns supposing that L is equal to that of the input transformer, viz. 20 μH, so:

$n= (L.l/\mu_o.\mu_r.A)^{\frac{1}{2}}= (20x10^{-6}x92x10^{-3}/4x\pi x10^{-7}x120x97,6x10^{-6})^{\frac{1}{2}}= 11,2$ turns

As we will see further on this number of turns must be even so we choose:

$n_{sec}= 12$ turns, by which L becomes 23 μH.

The next step is to calculate B_{max} for which we need V_{max}.
The latter can be calculated from the maximum output power and the impedance R_L of 50 Ohms. In this case the maximum power amounts to 300 Watts, corresponding to:

$V_{max}= (2.Po.R_L)^{\frac{1}{2}}= (2x300x50)^{\frac{1}{2}}= 173,2$ Volts across 50 Ohms.

So:

$B_{max}= V_{max}/ (\omega_{min}.A.n)= 173,2/ (2\pi x1,6x10^6x97,6x10^{-6}x12)= 1,47x10^{-2}T.$

The core loss can be calculated with the aid of Fig. 4.
Then we need the quantity $B_{max}.f$ being equal to:
$1,47x10^{-2}x1,6x10^6= 2,35x10^4$. From Fig. 4 we can see that:
$\mu_r.Rp/L= 2,9x10^{10}$ sec $^{-1}$, so:
$Rp= 2,9x10^{10}L/\mu_r= 2,9x10^{10}x23x10^{-6}/120= 5560$ Ohms.

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From this it can be concluded that the core loss is:
(50/5560) x 100% = 0,9% of 300 Watts or 2,7 Watts.

At the primary side the impedance is 2x6,25 = 12,5 Ohms what determines the number of turns to: $n_{prim} = n_{sec}\ (Rp/Rs)^{\frac{1}{2}} = 12\ (12{,}5/50)^{\frac{1}{2}} = 6$ turns.

The practical realization is shown on page 28.

The L_{Str} measured at the two outer frequencies with short-circuited primary (low ohmic side) is:

f(MHz)	Xs(Ohm)
1,6	2,037
28	34,99

2.7 H.F. COMPENSATION OF THE OUTPUT TRANSFORMER

The calculation is in main lines similar to the foregoing, so it shall be given briefly.

12,5 ⟶ 50 Ohms transformation

$X_{L\ max} = 2\pi \cdot f_{max} = 34{,}99$ Ohms

$f_{max} = 28$MHz.

$X_{L\ max}/R = 34{,}99/50 = 0{,}7 \therefore$ VSWR = 1,03

From the curve in ref. 6 it follows that:

$R/X_{C\ min} = 0{,}4 \quad X_{C\ min} = 50/0{,}4 = 125$ Ohms

$C_{sec} = 1/2\pi\, f_{max}\ X_{C\ min} = 1/2\pi \times 28 \times 10^6 \times 125 = 45{,}3$pF

$C_{prim} = T^2 \times C_{sec} = 2^2 \times 45{,}3 = 181{,}2$pF

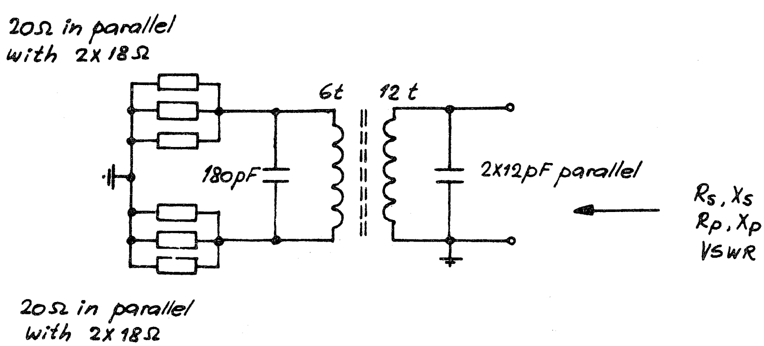

Fig. 10

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f(MHz)	Rs(Ohm)	Xs(Ohm)	Rp(Ohm)	Xp(Ohm)	VSWR
1,6	50,24	8,79	51,78	296	1,19
3,5	51,83	3,31	52,04	815	1,08
7	52,46	0,26	52,46	10407	1,05
14	54,00	0,69	54,01	4237	1,08
20	54,30	2,96	54,46	999	1,11
24	54,88	6,08	55,55	501	1,16
28	54,98	8,28	56,22	373	1,20

2.8 L.F. COMPENSATION OF THE OUTPUT TRANSFORMER

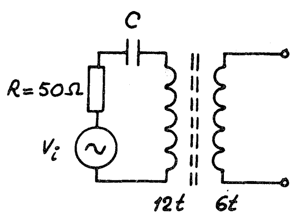

Fig. 11a

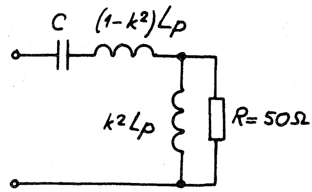

Fig. 11b

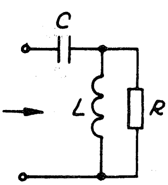

Fig. 11c

L_{prim} = 23 μH (calculated)

according to ref. 8.

$R/X_{L\ min} = X_{c\ max}/R$

$X_{L\ min}$ = 231 Ohms

$R/X_{L\ min}$ = 50/231 = 0,2165 ∴ VSWR = 1,04

$X_{c\ max}$ = 0,2165 x 50 = 10,8 Ohms.

$C = 1/2\pi \cdot f_{min} \cdot X_{c\ max} = 1/2\pi \times 1{,}6 \times 10^6 \times 10{,}8$ = 9,2nF

4,7nF parallel with 5,6nF was chosen because this particular core had a somewhat higher permeability.

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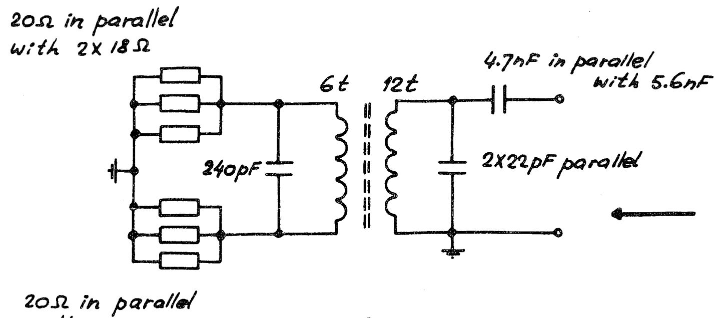
2.9 THE COMPENSATED OUTPUT TRANSFORMER



The primary compensation capacitor has been increased to 240pF to compensate for the inductance of the load resistors.

R_s, X_s
R_p, X_p
VSWR

Fig. 12

The table below shows the results:

f(MHz)	Rs(Ohm)	Xs(Ohm)	Rp(Ohm)	Xp(Ohm)	VSWR
1,6	50,49	1,63	50,55	1570	1,03
3,5	50,92	1,44	50,96	1798	1,03
7	50,92	1,59	50,97	1636	1,04
14	50,91	2,38	51,02	1092	1,05
20	50,77	3,10	50,96	834	1,07
24	50,11	4,02	50,44	628	1,08
28	50,13	3,71	50,41	681	1,08

As can be seen from the table the VSWR has not only been improved at low frequencies but also at high frequencies.

2.10 THE CENTRE TAPPED CHOKE COIL

According to the amplifier described in ECO 7114 (ref. 5) the output circuit contains a centre tapped choke coil, that in our case consists of a FXC rod instead of a tube of 4B1 material with windings made of twisted enamelled copper wire.

Because a good explanation of the behaviour has been given in the mentioned report the description has been omitted in this report.

The twisted line has a characteristic impedance of appr. 45 Ohms, a capacitance of 155pF/m, a coupling factor between the windings of 97,5% and an inductance of 1,44 μH typical for one winding.

For the normal range of operation the equivalent circuit of the choke coil consists of an inductance L_{CH} = 5,75 μH in parallel with a capacitance C_{CH} = 7pF.

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2.11 COMPENSATION MEASURES IN THE OUTPUT CIRCUIT

The total collector load is formed by the circuit shown in Fig. 13 in which Co is the transistor output capacitance and

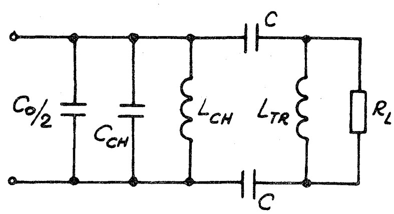

Fig. 13

C a d.c. blocking capacitor. The value of L_{TR} amounts to 5,75 μH being transformed from the secondary side.

According to ref. 8 L.F. compensation can be applied. In that case the circuit has to be divided into two parts. Fig. 14 shows the new situation, which the influence of C_o and C_{CH} has been neglected.

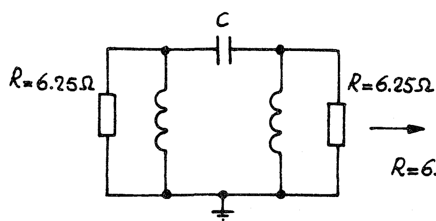

Fig. 14a

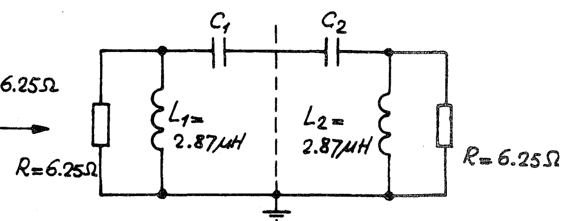

Fig. 14b

In this case it happens to be that $L_1 = L_2$, so that:

$C = C_1/2 = C_2/2$

$R/X_{L\ min} = X_{C1\ max}/R$

$X_{L\ min} = 2\pi . f_{min} . L_1 = 2\pi \times 1,6 \times 10^6 \times 2,87 \times 10^{-6} = 28,85$ Ohms.

$R/X_{L\ min} = 6,25/28,85 = 0,2165 \therefore$ VSWR= 1,04

$X_{C1\ max} = 0,2165 \times 6,25 = 1,352$ Ohm.

$C_1 = 1/2\pi . f_{min} . X_{C1\ max} = 1/2\pi \times 1,6 \times 10^6 \times 1,352 = 73,5$nF

So the d.c. blocking capacitors, giving L.F. compensation too, have to be: C= 73,5/2 = 36,7nF.

Because tne effective current through these capacitors is very high, viz. appr. 5A they have been composed of a number of capacitors in paralle In case of 3 capacitors the current is $1,63A_{eff.}$ being still too high to be handled by a polyester capacitor, because of too high tgδ. Because of the lack of a suitable type it had been tested with mentioned parallel connection. It operated without troubles. ../19

However the problem has been solved later on by using polystyrene capacitors (3x 11nF ± 5% in parallel, V= 125 Volts). This value is somewhat lower than the calculated one because in the design phase it was assumed that $L_1 = L_2 = 2{,}5\,\mu H$; based on performance there was no reason to change it afterwards.

Connecting 3 pieces in parallel keeps the total series-inductance low, what is required from the point of view of H.F. performance.

The H.F. compensation could be arranged as follows:

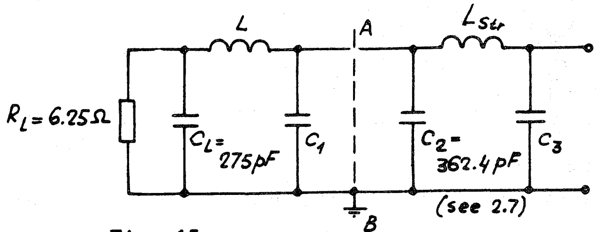

Fig. 15

In the right hand part of Fig. 15, the π section with C_2, L_{Str} and C_3 has an input resistance of 6,25 Ohms.

The R_L and C_L of resp. 6,25 Ohms and 275pF have to be matched to the 6,25 Ohms of the network. This could be done by introducing a second π network that consists of C_L, L and C_1.

However a simpler solution is possible (See Fig. 16) when the value of L is so that it compensates the capacitance C_L.

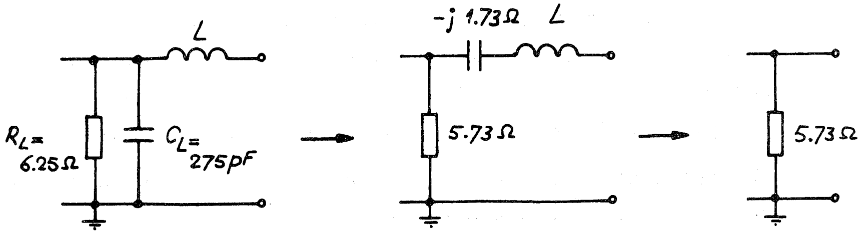

Fig. 16a   Fig. 16b   Fig. 16c

Transforming R_L and C_L to series components (Fig. 16b) they become 5.73 Ohms in series with -j 1,73 Ohm. For compensation the value of L at 28MHz has to be +j 1,73 Ohm or 9,8nH, a value being approximately present in the circuit (wiring, p.c.board). The somewhat decreased R_L = 5,73 Ohms for the highest operating frequency has been accepted, so C_1 appeared to be superfluous.

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2.12 <u>THE BIAS CIRCUIT</u>

The amplifier operates from a 50 Volts supply voltage, what has some consequences for the bias circuit.

Fig. 17 shows the applied circuit.

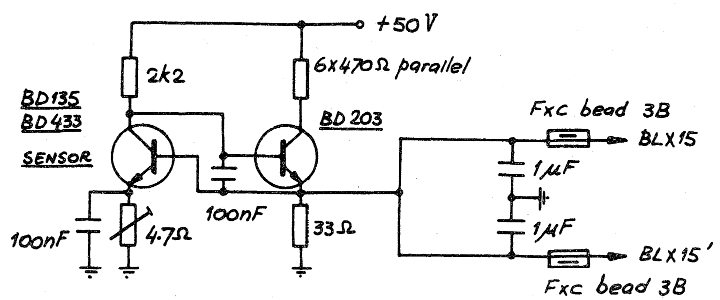

Fig. 17

The maximum current drain of the emitter follower depends on the d.c.-collector current of the BLX15s and their minimum h_{FE}. The transistor type can be supplied in 7 h_{FE} groups viz. BLX15A (I_B= 73,5 - 87,5mA for I_C= 1,4A and V_{CE}= 6 Volts)
to BLX15G (I_B= 27,5 - 31,5mA for I_C= 1,4A and V_{CE}= 6 Volts).
The minimum h_{FE} for the A type amounts to 15.

The collector current has its maximum around 20 - 28MHz. A value measured in a practical amplifier was 12,2A. This gives with the minimum h_{FE} of 15 a total base current of 815mA.
The current capability depends mainly on the value of the collector series resistor of TR2. The voltage across this resistor has a maximum of appr. 49 Volts if the transistor is in bottoming. From the above follo a value of 49/0,815 = 60 Ohms. In practice a some what higher value *) has been chosen (6 x 470 Ohms in parallel), having a maximum dissipation of appr. 30 watts, however this dissipation will always be lower.

The collector dissipation of TR2 has a maximum value at the half V_B point, namely appr. $25^2/78$ = 8 Watts.
An NPN silicon epitaxial base transistor BD203 can do the job (V_{CEO}= $60V_{max}$., $I_{C\ max}$= 8A, P_{tot}= max. 60Watts). It can be thermally connected to the main heatsink.

*) The practical minimum h_{FE} is higher than 15 due to the fact that this quantity increases as a function of V_{CE}, I_C and junction temperature.

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The bias voltage will be equal to the sum of the base-emitter voltage of TR1 and the voltage across its emitter resistor.
A requirement is that the V_{BE} must have such a value that it is below that needed to cause appr. 100mA collector current in each BLX15.
Adjustment of the quiescent current with a variable emitter resistor is then possible.

In the first instance a BD135 was applied, but due to spread in characteristics of that type we decided to change to a BD433 having a lower V_{BE} because of a larger crystal.
TR1 is thermally coupled to the heatsink of the r.f. transistors and as close to these devices as possible.
By doing so the bias will compensate the 2mV/°C drop of the V_{BE} of the BLX15s.

The power dissipation in TR1 is only 33m Watts. There are no problems with the V_{CEO} of that type, because it is clamped to two emitter base junctions so $V_{CE} \approx 1{,}5$ Volts.

The internal resistance of the circuit was measured. It appeared to be appr. 30m Ohms.

The capacitors of 100nF, 1 µF and the FXC beads prevent parasitic oscillations.

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3. THE PRACTICAL AMPLIFIER

Fig. 18 shows the circuit diagram of the complete amplifier.

Constructional details of the transformers have been given in Figs. 19, 20 and 21.
For a better symmetry both toroids have been mounted perpendicularly to the board.
From the practical point of view it was impossible to connect the centre ends of the collector choke windings to the same point. So two separate decoupling capacitors were applied.
Each point is decoupled by three capacitors of 100nF in parallel. This was done to obtain a low reactance for the lower frequencies and a small lead inductance for the higher frequencies.

The series inductance of the applied type is appr. 8nH for 2 x 2,5mm lead length. The decoupled points could not be interconnected directly. Insertion of two modified FXC chokes was necessary to shift the resonant frequency to below the lowest operating frequency.

The components were mounted on a p.c.board of epoxyglass. The lower sheet of the double clad board functions as a ground plane. Interconnections of some upper parts with the ground plane were made with 2mm tubular rivets, being soldered to the tracks to be sure of contact.

In the laboratory test sets the BLX15s housing in plastic SOT-55 envelopes were not soldered in the circuit, but their leads were pressed on contact plates by small screws through the holes in the leads.

The bias network is on a small p.c.board. It was screwed against the heatsink.
Fig. 22, 23, 24 and 25 show the p.c. boards with situation of the components.

The cross-neutralising network has to be as short as possible; small series resistors have been inserted to prevent oscillations.

The laboratory models have been water-cooled, the bias transistors were screwed against the heatsink via mica washers.

It is advisable to apply a thermally conducting paste for the BLX15s.

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4. MEASUREMENTS

4.1 GENERAL

Three amplifiers were built and measured. All the results of measurements will be given.
The measurements were done under the nominal conditions:
Battery voltage: V_B= 50 Volts.
Load and source impedance: R_L and R_s = 50 Ohms.
Ambient temperature: T_{amb} appr. 25°C.

4.2 INTERMODULATION DISTORTION

The I.M.D. versus output power is measured with a two tone signal (p.q) at the following frequencies: 1,6; 3,5; 7; 14; 20 and 28MHz.
This signal is made as follows: The output signals of two X-tal controlled oscillators with a difference of 1kHz, switchable in pairs at above frequencies, have been separately amplified in two linear broad-band amplifiers. These signals have been combined in a hybrid and supplied to a driver amplifier operating in class A. The latter must be able to supply sufficient power to the input of the final.

Two class A broadband amplifiers intended for driver applications have been considered:

1) A single stage amplifier with one BLX13, delivering an output power of 8 Watts P.E.P. for an I.M.D. $\leqslant$ -40dB. The gain = 16,8 $\pm$ 0,2dB. The amplifier has been described in C.A.B. report ECO 7113 (Ref. 9).

2) A single stage amplifier with two BLX13s in push-pull with an output power of 14 Watts P.E.P. for an I.M.D. $\leqslant$ -40dB. The gain = 18,7 $\pm$ 0,15dB. A description is in report ECO 7201 (Ref. 10).

The first tests were performed with the smaller driver, appearing barely enough for finals with a lower power gain. So the second one, being used for later experiments, is recommended.

The oscillator part of the two-tone set-up also contains X-tal controlled oscillators presenting the local oscillator signals to the spectrum analyzer SINGER MF5.

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Fig. 26 shows the block diagram.

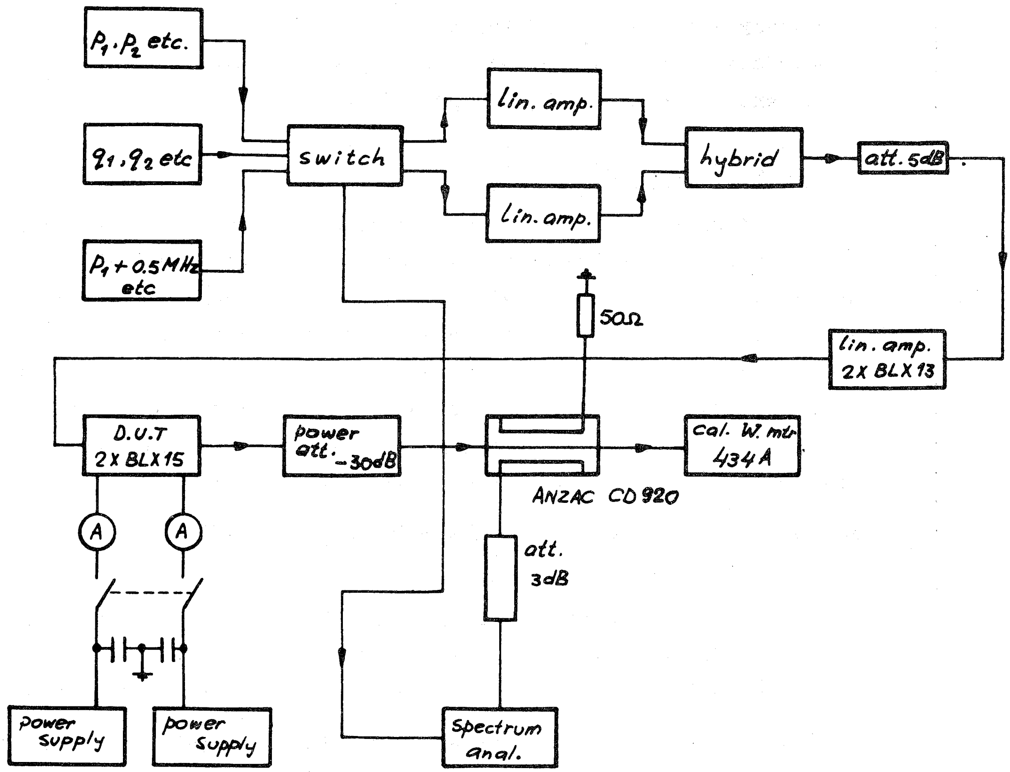

Fig. 26

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4.3 INPUT VSWR, EFFICIENCY AND POWER GAIN

These single tone measurements have been done with the set-up of Fig. 27.

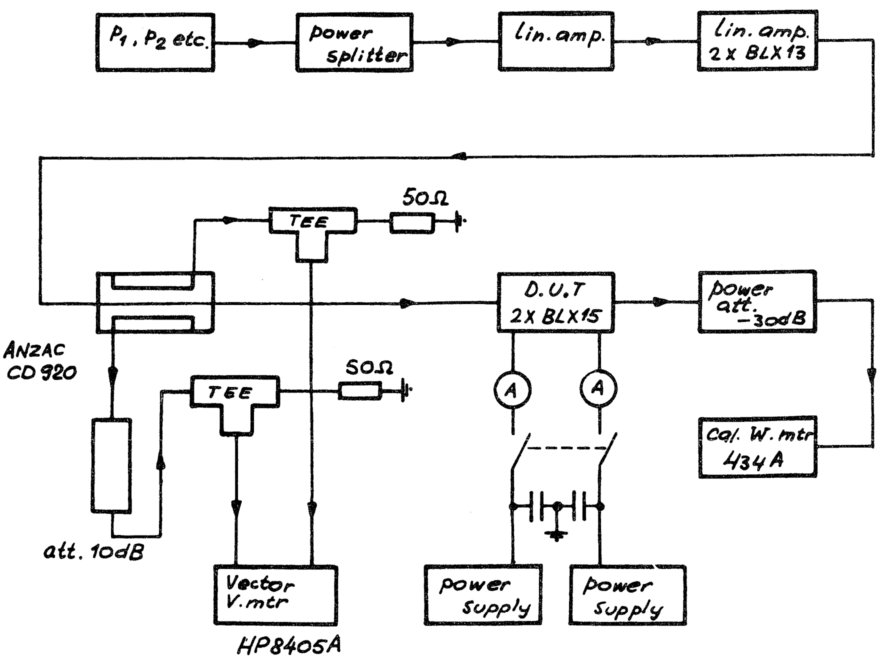

Fig. 27

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4.4 MEASURED RESULTS (I.M.D.) PROTOTYPE AMPLIFIER

The intermodulation is measured at output powers of 30, 100, 200 and 300 Watts P.E.P. In case of unequal d_3 and d_5 components the worst result is given. (d_3 and d_5 in dB)

	P_o = 300 Watts		P_o = 200 Watts		P_o = 100 Watts		P_o = 30 Watts	
f(MHz)	d_3	d_5	d_3	d_5	d_3	d_5	d_3	d_5
1,6	-38	-42	-48	-41	-37	-34	-33	-40
3,5	-42	-41	-38	-42	-37	-35	-33	-41
7	-45	-41	-36	-43	-52	-34	-39	-47
14	-31	-44	-32	-40	-36	-33	-40	-43
20	-35	-38	-34	-38	-32	-35	-36	-34
28	-30	-38	-42	-41	-37	-37	-34	-42

Figs. 28a and 28b show the d_3 versus frequency.
In this case the driver with one BLX13 (Ref. 9) has been used.

4.5 MEASURED RESULTS (INPUT VSWR, EFFICIENCY AND GAIN) OF PROTOTYPE AMPLIFIER

Typical results are given for an output power of 300 Watts.

f(MHz)	I_{C1}(A)	I_{C2}(A)	eff(%)	P_{if}(W)	gain (dB)	VSWR
1,6	5,35	4,80	59,2	5,58	17,30	1,03
3,5	5,12	4,90	58,8	6,08	16,93	1,04
7	5,25	5,32	56,8	6,18	16,86	1,09
14	6,05	5,95	50,0	6,82	16,44	1,07
20	6,30	5,90	49,2	6,28	16,79	1,17
28	6,45	5,63	49,7	7,06	16,28	1,06

For curves see Fig. 29.

I_{Css} = 2x 120mA

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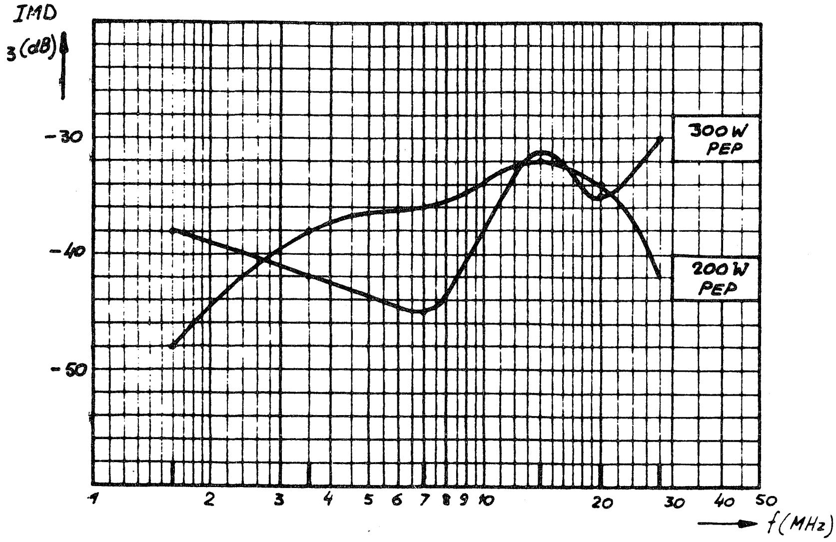


Fig. 28a

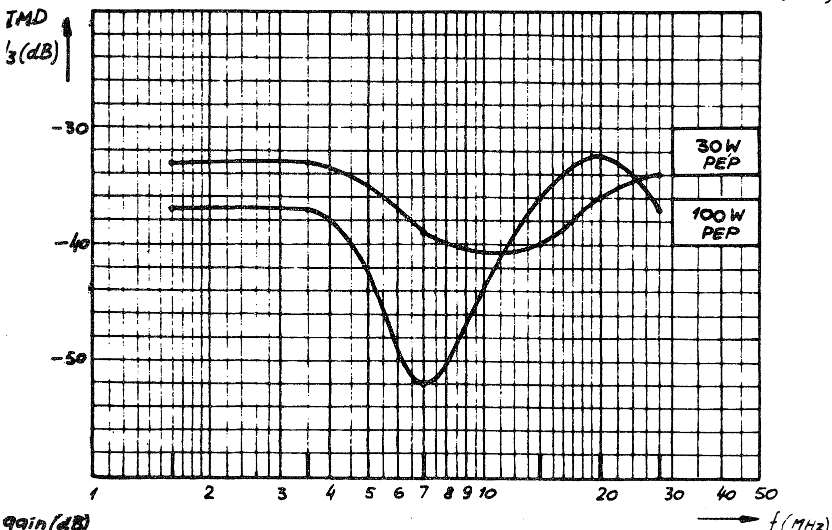


Fig. 28b

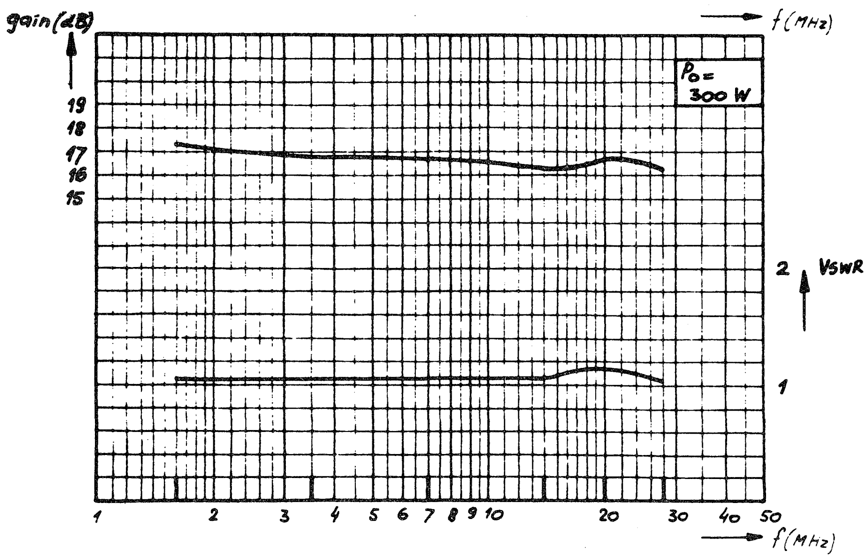


Fig. 29

../28

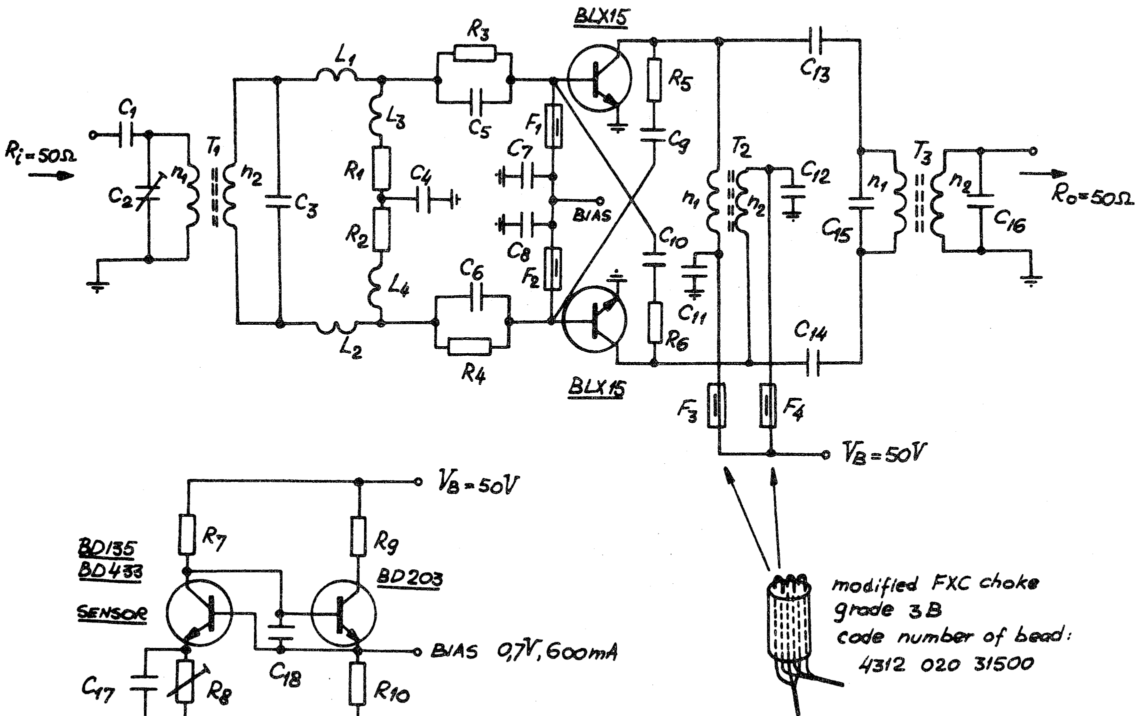


Fig. 18

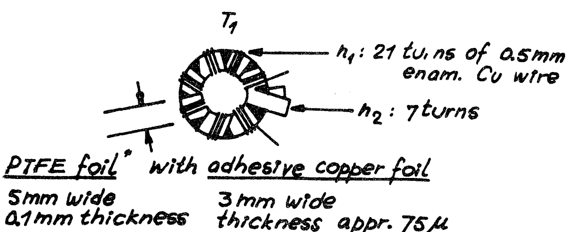


Fig. 19

Core: Fxc toroid, grade 4C6, 14x9x5 mm^3
code number: 4322 020 91020

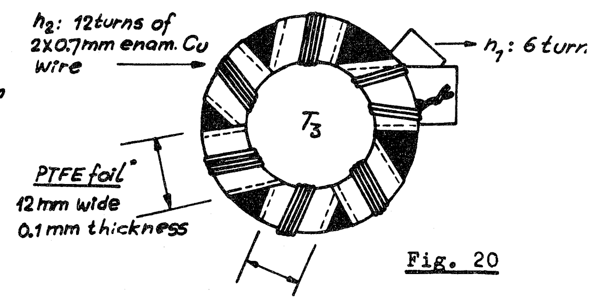
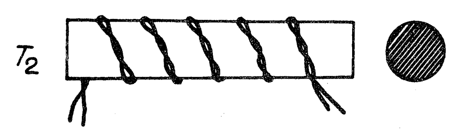


Fig. 20

adhesive copper foil
10 mm wide, thickness appr. 75μ

* PTFE foil on the outside

Core: FXC toroid, grade 4C6, 36 x 23 x 15 mm^3
code number: 4322 020 91090



n_1, n_2: 5 turns of 2x1.0mm twisted enam. Cu wire
wound on a Fxc rod , grade 4B1
Dimensions (Dxℓ): 10x 50 mm^2
Part of antenna rod: 3122 104 91250

Fig. 21

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5. <u>COMPONENTS</u>

R_1, R_2	= parallel connection of: 3x 10 Ohms, carbon ± 5%, CR68 style	2322	214	13109
R_3, R_4	= parallel connection of: 2 x 18 Ohms, carbon ± 5%, CR25 style	2322	101	33189
R_5, R_6	= parallel connection of: 3 x 10 Ohms, carbon ± 5%, CR37 style	2322	212	13109
R_7	= 2,2k Ohms, carbon ± 5%, CR52 style	2322	101	63222
R_8	= 4,7 Ohms wire-wound trimming pot.meter, 2 Watts	2322	011	02478
R_9	= parallel connection of: 6 x 470 Ohms, enamelled wire-wound ± 5% - 5,5 watts each	2322	320	32471
R_{10}	= 33 Ohms, carbon ± 5%, CR25 style	2322	101	33339
C_1	= parallel connection of: 2x4,7 nF polyester ± 10%	2222	342	45473
C_2	= 4-100pF film dielectric trimmer	2222	809	07015
C_3	= parallel connection of: 2x560pF miniature polystyrene ± 1%	2222	425	45601
C_4, C_{17}, C_{18}	= 100nF polyester ± 10%	2222	342	45104
C_5, C_6	= parallel connection of: 3x1200pF polyester ± 10%	2222	342	45124
C_7, C_8	= 1 μF moulded metallised polyester ± 10%	2222	344	21105
C_9, C_{10}	= 82pF ceramic	2222	555	56829
C_{11},C_{12}	= parallel connection of: 3x100nF polyester ± 10%	2222	342	45104
C_{13},C_{14}	= parallel connection of: 2x10nF polyester and	2222	342	45103
	1x12nF polyester	2222	342	45123

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C_{15}	= 180pF ceramic	2222	555	56181
C_{16}	= parallel connection of: 2x22pF ceramic	2222	555	56229
T_1	= see Fig. 19			
T_2	= see Fig. 21			
T_3	= see Fig. 20			
L_1, L_2	= 13nH; 0,5 turn of 1mm copper wire D_{int} = 6mm, with 2 x 6mm leads			
L_3, L_4	= 14, 3nH; 0,5 turn of 1mm copper wire, D_{int} = 6mm, with 2 x 7mm leads			
F_1, F_2	= ferroxcube choke, grade 3B	4312	020	36640
F_3, F_4	= see Fig. 18			

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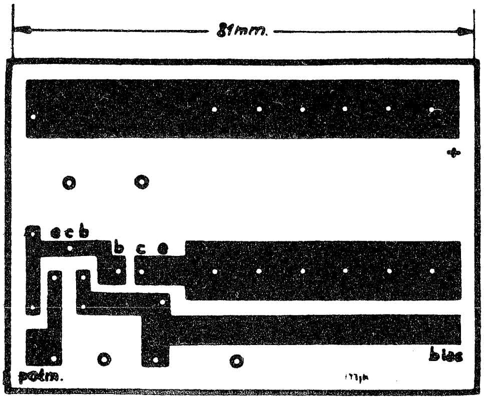
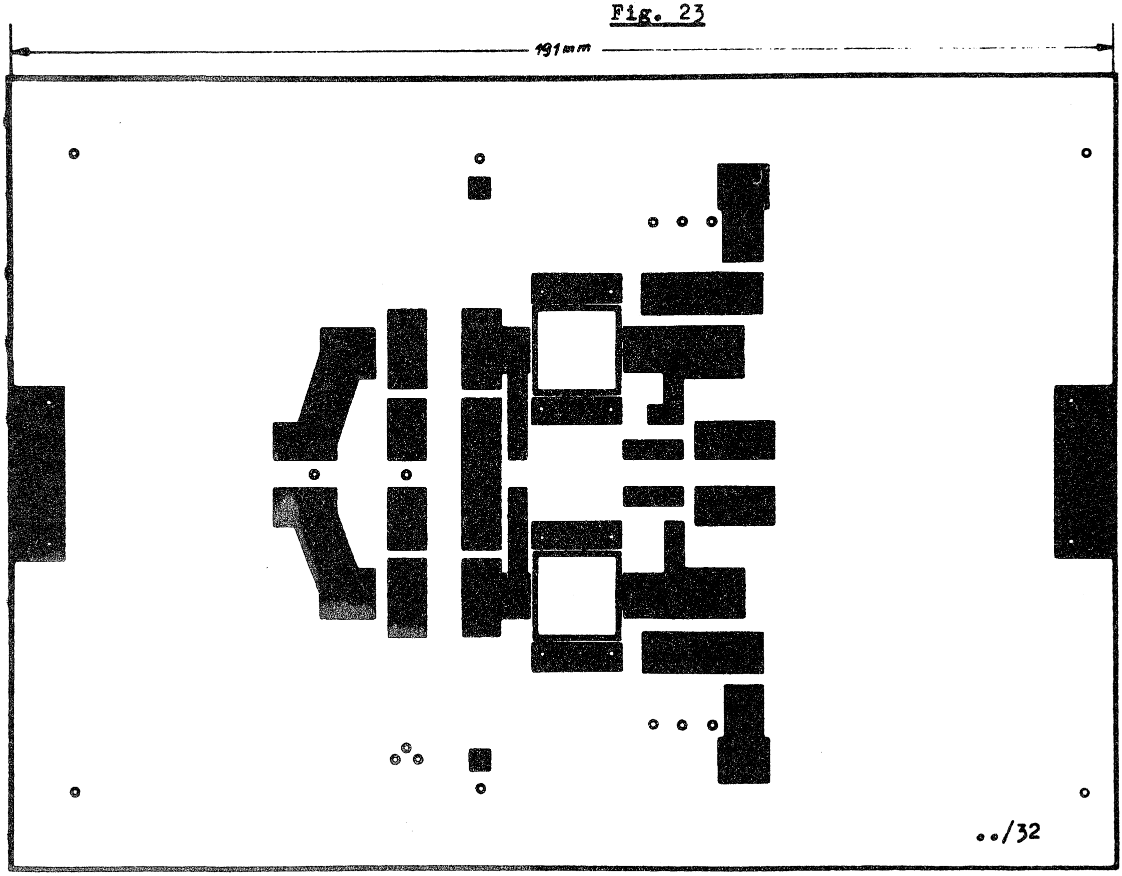


Fig. 22

Fig. 23



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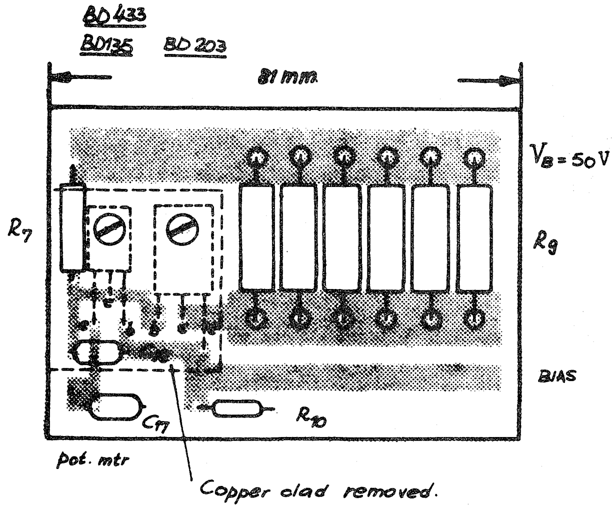
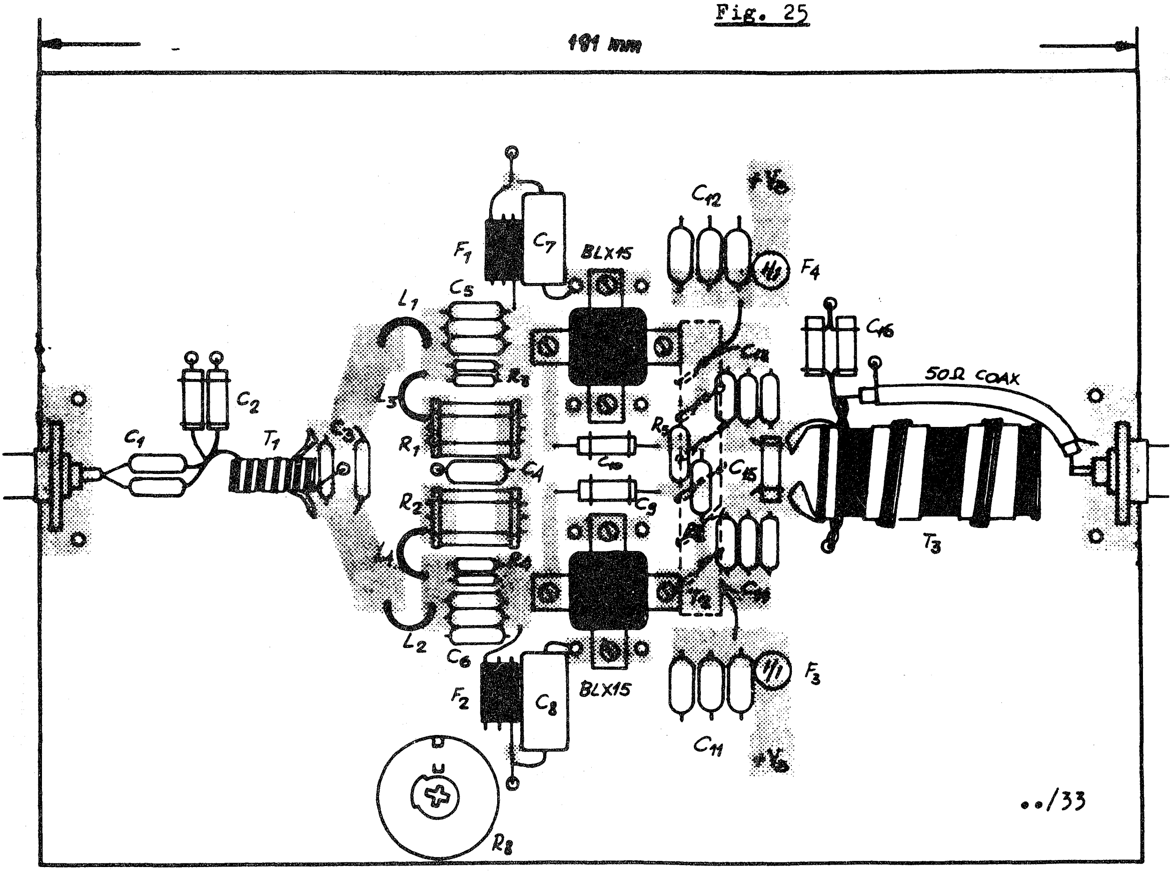


Fig. 24

Fig. 25



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6. MODIFICATIONS AND EXPERIENCES WITH RELEASED TRANSISTORS

In the preceding part of this report a description has been given of an amplifier design at the time of the appearance of the first development samples of the BLX15 (development number 402BLY).

Later on, around the period that the transistor was released, some modified amplifiers have been built in which measurements have been made with **matched** pairs of the typical BLX15.

These modifications mainly relate to changes of the p.c.board and types of components.
The p.c.board, applied in Fig. 23 came from an earlier design (Ref. 5) with 2 pieces BLX14 for 80 - 100 Watts P.E.P.

The new board is given in Fig. 26, whilst the modified lay-out is shown in Fig. 27.

Comparing both lay-outs shows that modifications exist mainly in widening or moving of tracks to obtain more room for the mounting of components.

In the mean time it also appeared that an often applied type of tubular ceramic capacitor (C_9, C_{10}, C_{16}) was cancelled and had to be replaced by the miniature ceramic plate type of class 1B (temperature compensating type with rated voltage of 500 Volts).

Chapter 3 mentions that the cross-neutralizing network has to be as short as possible. Special attention has been paid to that point.

In 2.11 it has been said that the effective current through coupling capacitors C_{13} and C_{14} is too high and because of lack of a suitable type it was tested with the mentioned parallel connection. The problem has been solved by using 3 polystyrene capacitors ("minipoco" type) in parallel, each having a value of 11nF $\pm$ 5%, V= 125 Volts.

Because the sizes are greater the tracks on the p.c.board have been changed. In that way a low series inductance could be maintained.

As a sensor in the bias circuit the BD433 has been chosen finally, whilst the value of C_7, C_8 has been increased to 3,3 μF to prevent parasitic oscillations.

In the first tests the value of C_2 was chosen equal to the calculated one. Soon it appeared that a variable type could better be applied, the value being defined from a tuning procedure.

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The following list gives a survey of the changed components, whilst the diagrams of Figs. 18 to 21 have been maintained.

6.1 CHANGED COMPONENTS

C_9, C_{10}	= miniature ceramic plate capacitor 500V d.c.	2222	650	58829
C_{13}, C_{14}	= parallel connection of:			
	3 x 11nF polystyrene ± 5%	2222	436	21103
C_{16}	= parallel connection of:			
	2 x 22pF 500V d.c.			
	miniature ceramic plate capacitor	2222	650	10229

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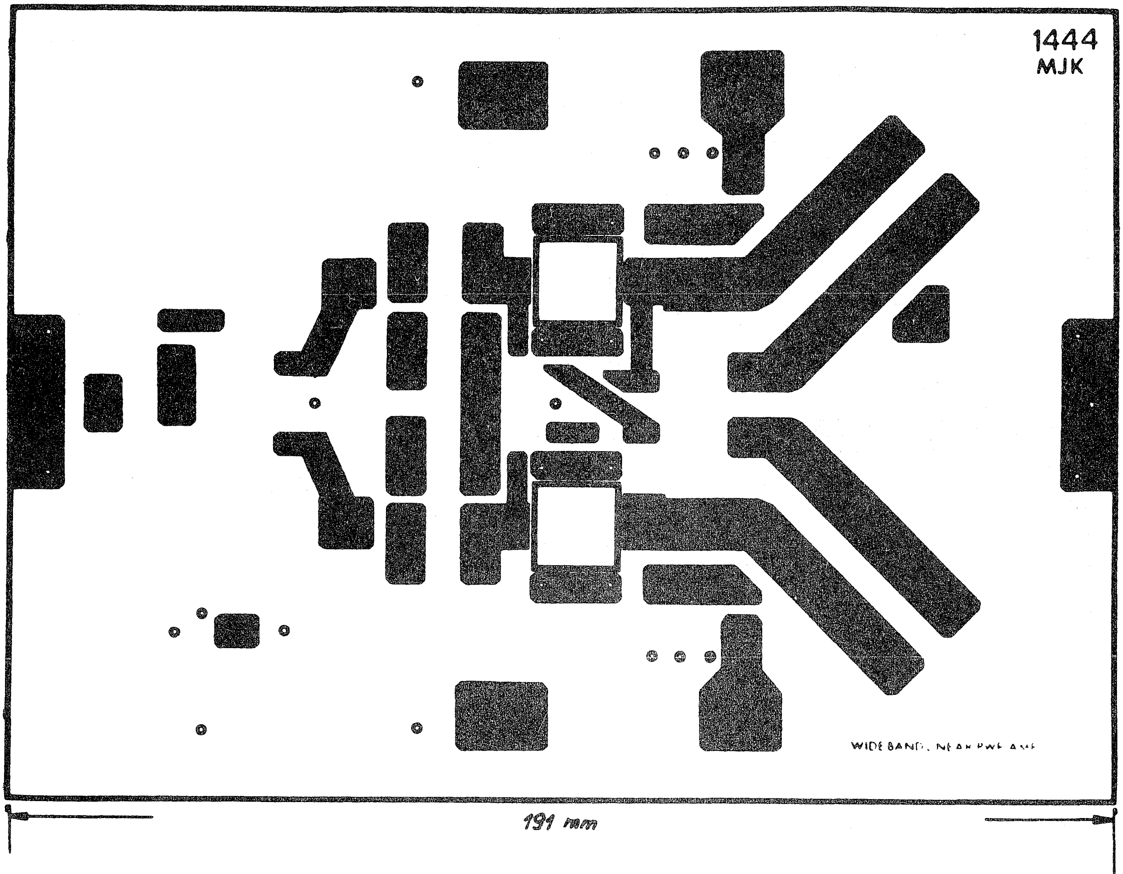

Modified printed circuit board

Fig. 26

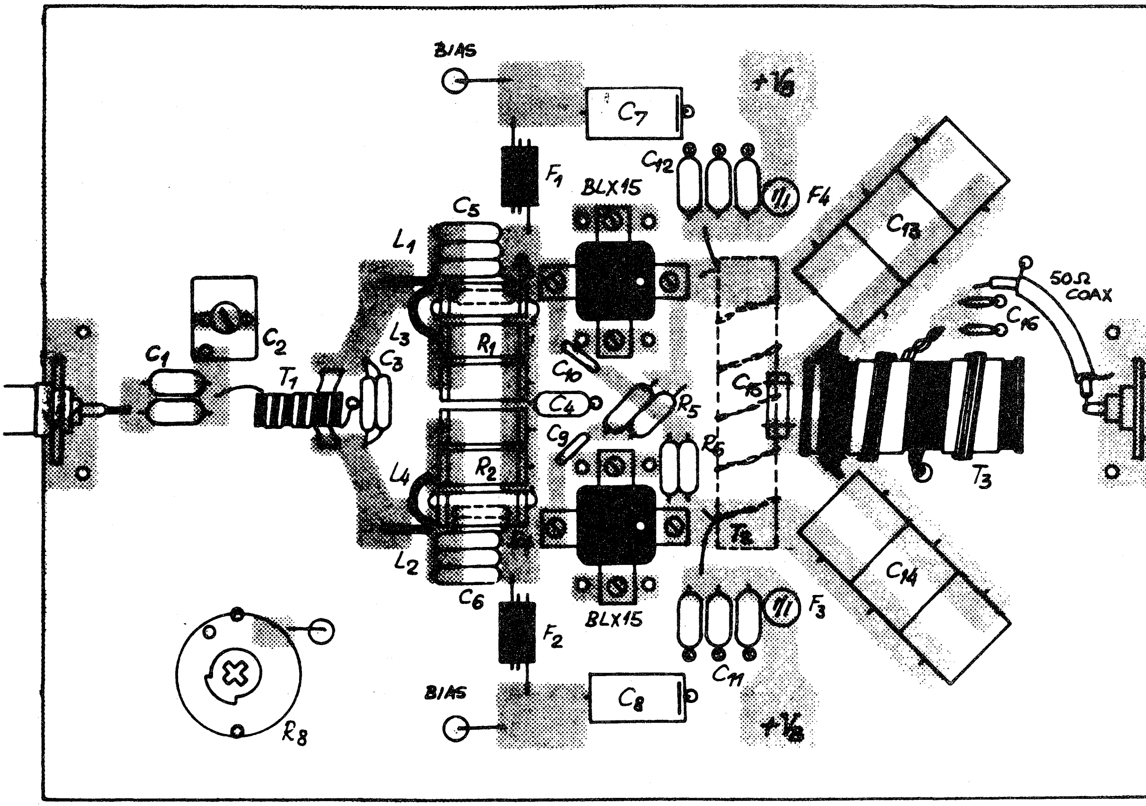
BIAS
C7
+VB
C12
F1
BLX15
F4
C13
C5
L1
50Ω
COAX
C2
C1
L3
R1
C16
T1
C3
C10
C15
C4
R5
C9
R6
T3
L4
R2
T2
L2
C14
C6
BLX15
F3
F2
C11
BIAS
C8
+VB
R8

Fig. 27

7. ALIGNMENT PROCEDURE

a) Insert the amplifier in the test circuit of Fig. 27 and **adjust the** total collector quiescent current at approx. 200mA for an ambient temperature of approx. 25°C.

b) Drive the amplifier up to a single tone output of 300 Watts at 20MHz and **tune** C_2 for minimum input reflection.
Note the value of the VSWR.

c) Repeat the measurement at 28MHz without detuning C_2.
Note the value of the VSWR.

d) Drive the amplifier up to a single tone output of 300 Watts at 28MHz and retune C_2 for minimum input reflection.
Note the value of the VSWR.

e) Repeat the measurement at 20MHz without detuning C_2.
Note the value of the VSWR.

f) Choose that alignment of C_2 that shows the smallest overall VSWR, what may result in repeating measurement b.

g) Measure the input VSWR, efficiency and power gain versus frequency as in 4.5.

h) Insert the amplifier in the test circuit of Fig. 26.

i) Measure the d_3 and d_5 for P_o = 30 Watts P.E.P. at 28MHz.

j) Turn R8 (I_{Css}) for the lowest IMD. The distortion will show a dip, as can be seen from a practical case in Fig. 28 in which that measurement has been done point by point.
Maintain the value of I_{Css} further on.
The VSWR, power gain and efficiency will hardly be influenced by the change of I_{Css}.

k) **Continue measurements as performed in chapter 4.4.**

8. RESULTS WITH TYPICAL RELEASED TRANSISTORS

Measurements have been done on the prototype with typical released transistors and on two modified circuits (1 and 2).

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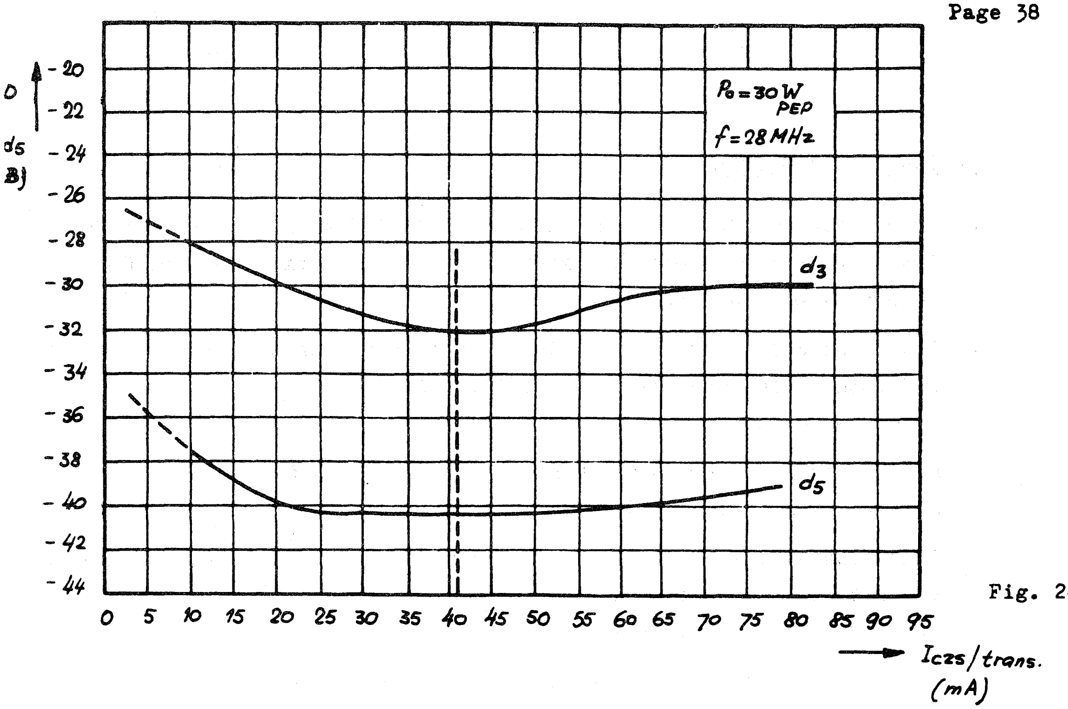

Fig. 28

8.1 PROTOTYPE WITH TYPICAL TRANSISTORS

Typical results for P_o = 300 Watts and V_B = 50 Volts.

f(MHz)	I_{C1}(A)	I_{C2}(A)	eff.(%)	gain (dB)	VSWR
1,6	4,64	4,88	63,0	15,24	1,01
3,5	4,64	5,05	61,9	14,95	1,08
7	4,74	5,23	60,2	14,88	1,15
14	5,40	6,40	50,8	14,10	1,14
20	5,75	6,10	50,6	14,77	1,22
28	5,75	5,56	53,0	15,24	1,35

Curves in Fig. 29.

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f(MHz)	P_o = 300 Watts d_3	d_5	eff (%)^{zt}	P_o = 200 Watts d_3	d_5	P_o = 100 Watts d_3	d_5	P_o = 30 Watts d_3	d_5
1,6	-40	-48	47,7	-42	-43	-38	-40	-36	-41
3,5	-43	-44	47,6	-48	-43	-39	-41	-33	-45
7	-42	-45	47,6	-50	-48	-41	-48	-33	-47
14	-31	-60	39,5	-35	-55	-38	-52	-34	-45
20	-33	-43	37,5	-35	-43	-35	-42	-33	-43
28	-31	-35	40,0	-40	-39	-33	-39	-31	-43

I_{Czs} = 2 x 41mA

Curves in Figs. 30a and 30b.

8.2 AMPLIFIER 1 WITH TYPICAL TRANSISTORS

Typical results for P_o = 300 Watts and V_B = 50 Volts.

f(MHz)	I_{C1}(A)	I_{C2}(A)	eff(%)	gain (dB)	VSWR
1,6	5,43	5,75	53,7	15,95	1,04
3,5	4,75	5,07	60,9	16,25	1,02
7	4,97	5,15	59,4	16,03	1,07
14	5,84	5,65	52,2	15,50	1,03
20	5,82	6,16	50,1	16,30	1,22
28	5,35	6,30	51,5	16,85	1,33

Curves in Fig. 31.

f(MHz)	P_o = 300 Watts d_3	d_5	P_o = 200 Watts d_3	d_5	P_o = 100 Watts d_3	d_5	P_o = 30 Watts d_3	d_5
1,6	-31	-40	-33	-38	-31	-36	-30	-38
3,5	-40	-40	-37	-38	-32	-37	-28	-38
7	-43	-43	-40	-43	-33	-41	-28	-40
14	-32	-42	-33	-45	-33	-46	-30	-43
20	-36	-40	-35	-41	-33	-41	-30	-43
28	-38	-41	-38	-42	-34	-42	-30	-41

I_{Czs} = 2 x 100mA

Curves in Figs. 32a and 32b.

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8.3 AMPLIFIER 2 WITH TYPICAL TRANSISTORS

Typical results for P_o = 300 Watts and V_B = 50 Volts.

f(MHz)	I_{C1}(A)	I_{C2}(A)	eff.(%)	gain (dB)	VSWR
1,6	4,65	5,15	61,2	16,35	1,07
3,5	4,65	5,60	58,6	16,07	1,03
7	4,80	5,10	60,6	15,74	1,09
14	5,60	5,74	52,9	15,00	1,08
20	5,85	6,00	50,6	15,53	1,10
28	5,50	6,30	50,8	15,78	1,25

Curves in Fig. 33

	P_o = 300 Watts		P_o = 200 Watts		P_o = 100 Watts		P_o = 30 Watts	
f(MHz)	d_3	d_5	d_3	d_5	d_3	d_5	d_3	d_5
1,6	-33	-41	-33	-38	-32	-36	-30	-36
3,5	-37	-42	-36	-39	-33	-37	-30	-38
7	-36	-53	-43	-47	-43	-42	-36	-41
14	-32	-45	-33	-42	-36	-41	-38	-41
20	-36	-40	-34	-38	-33	-37	-32	-37
28	-32	-43	-35	-40	-33	-39	-33	-40

I_{Css} = 2 x 140mA

Curves in Figs. 34a and 34b.

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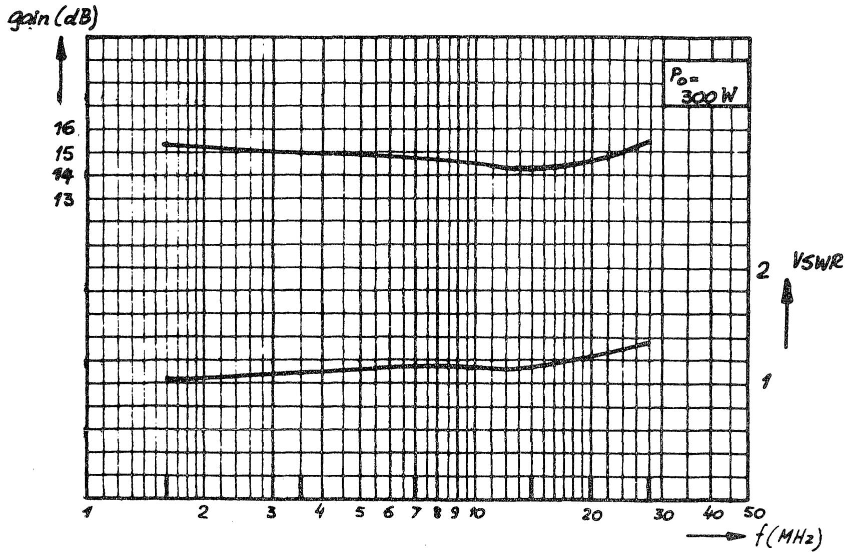

Fig. 29

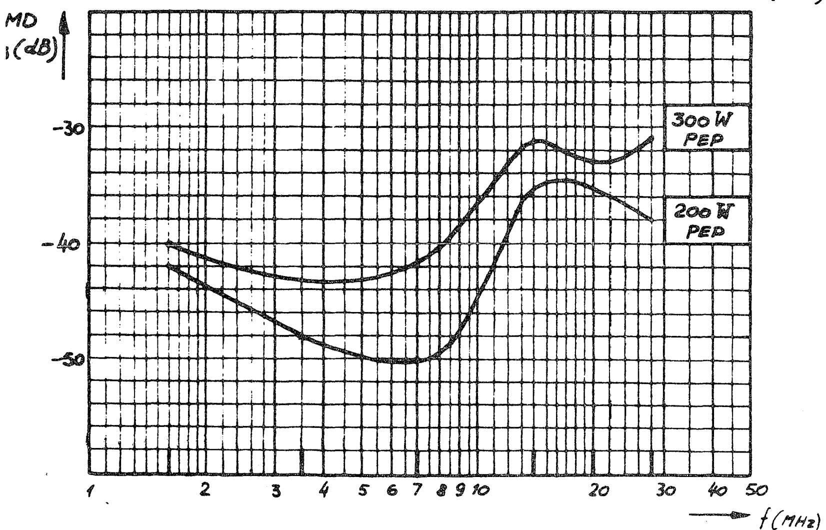

Fig. 30a

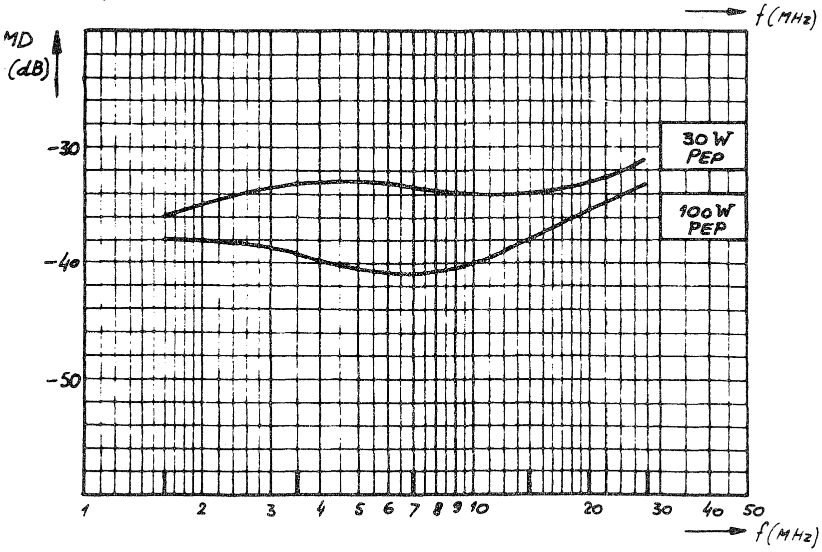

Fig. 30b

../42

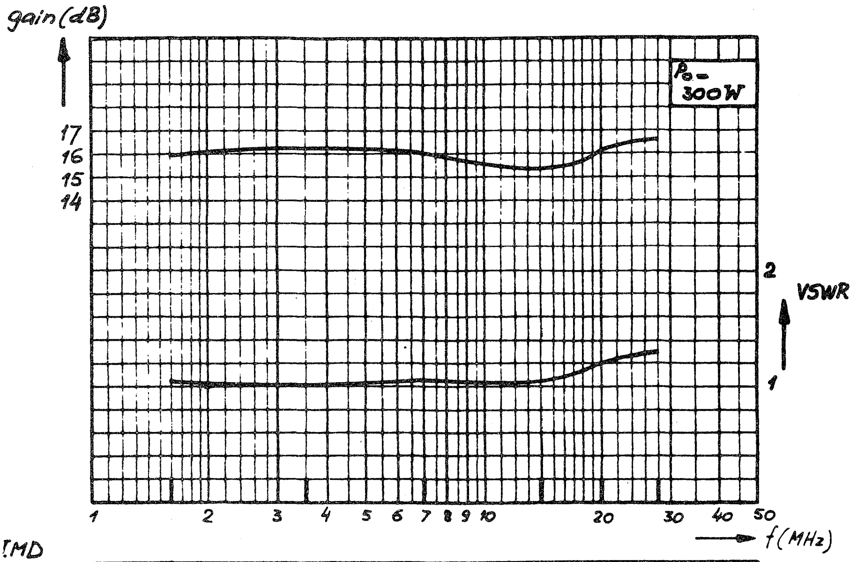

Fig. 31

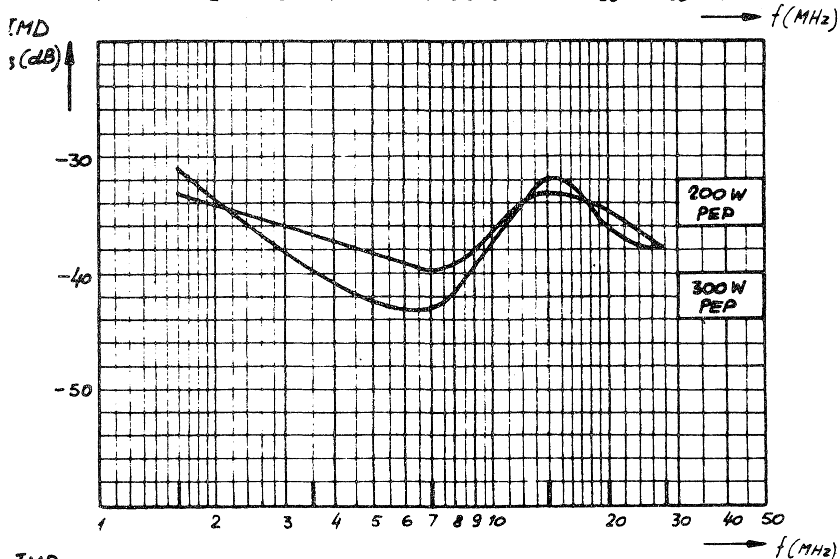

Fig. 32a

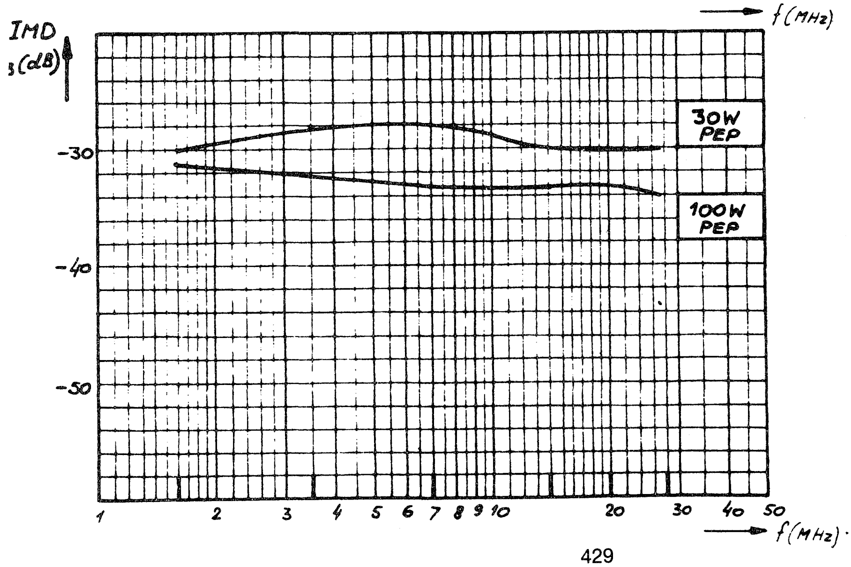

Fig. 32b

../43

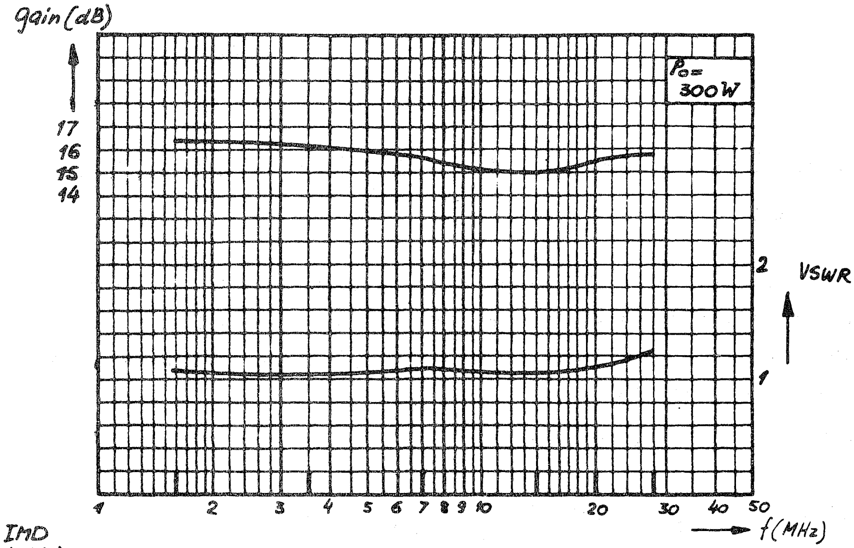

Fig. 33

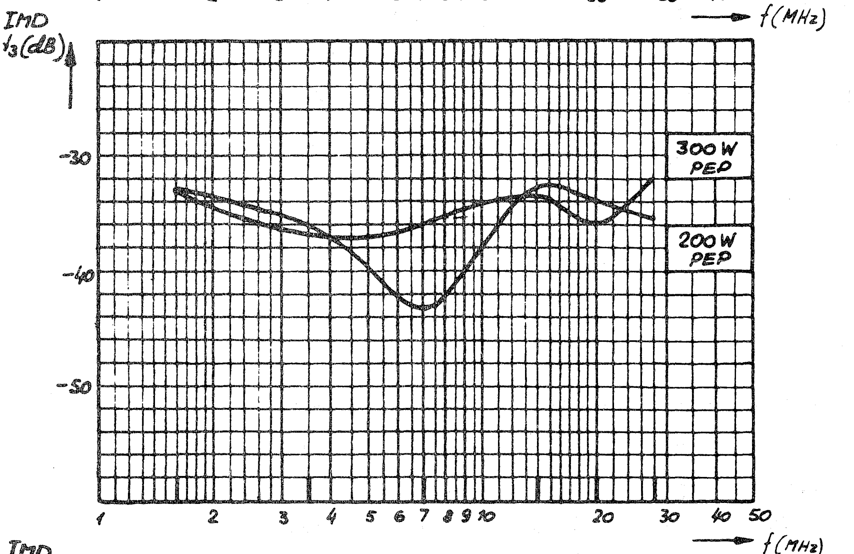

Fig. 34a

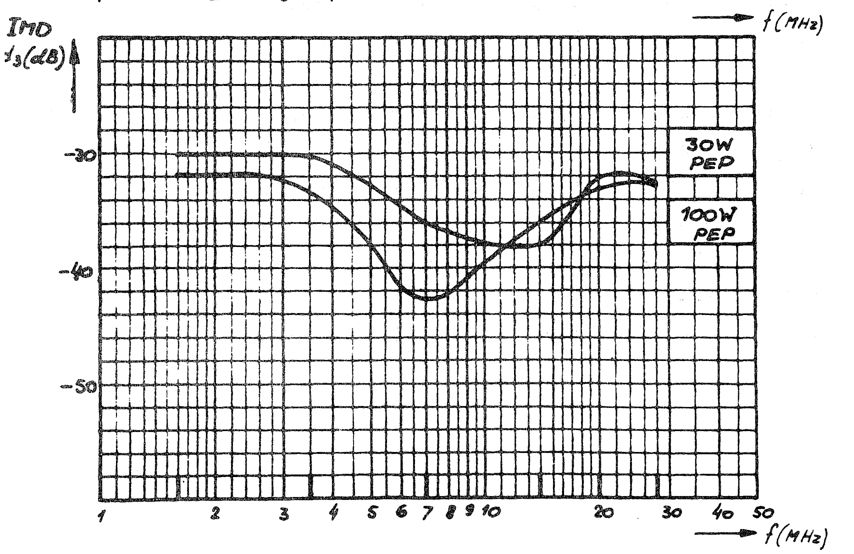

Fig. 34b

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9. COMPARISON OF FINAL RESULTS WITH COMPUTER CALCULATIONS BASED ON MOST RECENT PARAMETERS

Although it has been shown from the measuring results with released transistors that they come up to the expectations, the computer calculations, according to the previous description have been repeated with the values of the final product.

The table below gives the input data of the released type:

f_T = 270MHz	C_{be} = 2700pF	P_{load} = 150 Watts
h_{fe} = 30	C_{cbi} = 44pF	V_c = 45,5 Volts
R_b = 0,10 Ohm	C_{cbo} = 72pF	V_{bo} = 0,7 Volt
R_e = 0,11 Ohm	C_{ce} = 69pF	R_{bo} = 0,2 Ohm
R_c = 0,21 Ohm	C_s = 3,5pF	R_{be} = 10^{12} Ohms
L_b = 1,7nH		R_{bc} = 10^{12} Ohms
L_e = 1nH		C_n = 82pF
L_c = 1,5nH		

The next table shows the results on the equivalent circuit.

f (MHz)	R_L (Ohm)	C_L (pF)	R_i (Ohm)	X_i (Ohm)	G (dB)
1,6	6,25	-247,5	6,15	-2,02	29,00
2,5	6,25	-247,5	5,32	-2,64	28,88
3,5	6,25	-247,5	4,41	-2,93	28,69
5	6,25	-247,6	3,35	-2,89	28,31
7	6,25	-247,6	2,46	-2,53	27,68
10	6,24	-247,7	1,79	-1,97	26,59
14	6,23	-248,0	1,40	-1,42	25,06
20	6,20	-248,4	1,17	-0,89	22,93
24	6,17	-248,8	1,10	-0,64	21,67
28	6,15	-249,3	1,06	-0,44	20,54

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Calculation of the network components delivers the following values:

C_1 = 1118,2pF, L_2 = 13,3nH, R_3 = 3,38 Ohms
C_5 = 3489,1pF, L_3 = 21,4nH, R_4 = 9,19 Ohms

Now the same component values as in the prototype have been taken and the VSWR and gain versus frequency have been calculated.

The table below gives the output data of the computer in short form.

f (MHz)	R_{is} (Ohm)	X_{is} (Ohm)	VSWR (-)	gain (db)
1,6	2,76	0,01	1,01	17,75
2,5	2,77	0,01	1,01	17,64
3,5	2,78	0,01	1,00	17,52
5	2,79	0,00	1,01	17,39
7	2,82	-0,02	1,02	17,32
10	2,85	-0,09	1,04	17,38
14	2,80	-0,21	1,08	17,60
20	2,56	-0,17	1,11	17,97
24	2,50	0,12	1,12	18,09
28	2,76	0,60	1,24	18,00

It may be concluded from above results (gain = 17,7 ± 0,4dB, and $VSWR_{max.}$ = 1,24) that the amplifier circuit with components from original calculations can be maintained without much influence on the final results.

10. ACKNOWLEDGEMENT

The author wishes to express his thanks to Peter Coppens, who assisted in constructing both modified amplifiers and doing measurements.

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MITCHAM

UNCLASSIFIED REPORT

MC07404

A 1.0kW P.E.P. LINEAR POWER AMPLIFIER FROM F = 1.6MHz TO 30MHz USING BLX15 TRANSISTORS

BY

J. LING

AUTHOR:

J. LING

APPROVED:

P.G. GILES

DATE: 1974 JUL 09

JL/PGG/CAM
74-08-09

MC07404

A 1.0kW P.E.P. LINEAR POWER AMPLIFIER FROM F = 1.6MHz TO 30MHz USING BLX15 TRANSISTORS

ABSTRACT

A 1.0kW linear amplifier is described consisting of four wideband linear power amplifiers operating over the band f = 1.6MHz to f = 30MHz each with 2 BLX15 transistors in push-pull using a hybrid coupling system to provide 1.0kW p.e.p. output with intermodulation products better than -30dB when driven with a two-tone signal.

A single amplifier provides 300 Watts p.e.p. with intermodulation products better than -26dB and the reduction to 1.0kW from 1.2kW with four coupled amplifiers allows for the differences in performance of the individual amplifiers due to transistor spreads, component tolerances, insertion loss and variations of matching of the hybrid coupling system with frequency.

The measurements show that, with the combination of four units each using BLX15 transistors from the same h_{fe} grouping, 1.0kW p.e.p. can be developed in a wideband load with the intermodulation products at -30dB together with an overdriving capability of 1.2kW p.e.p. at -26dB.

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MC07404

A 1.0kW P.E.P. LINEAR POWER AMPLIFIER FROM F = 1.6MHz TO 30MHz USING BLX15 TRANSISTORS

1. INTRODUCTION

A linear power amplifier has been built consisting of four 300 Watt p.e.p. amplifiers that have been combined using a hybrid coupling system.

Each unit is equipped with a pair of BLX15 transistors operating in push-pull, taken from the same h_{fe} selection group. It should be noted that BLX15 transistors are supplied in seven h_{fe} groupings, A to G, covering the total range from 15 to 50 (in steps of 5).

The measurements shown have been made with transistors from group D.

2. CIRCUIT DESCRIPTION

The circuit for the individual amplifiers is shown in Figure 1 and a components list is given in Figure 2. The amplifiers were constructed on 2oz. double sided copper clad circuit board 191 x 135mm, 1.5mm thickness to the layout shown in Figure 3. Each amplifier was cooled using a copper block heatsink, which, in practice, can have fins for air cooling or be drilled for water cooling.

Four amplifiers were constructed and the performance spreads of the individual amplifiers together with the performance of the four units coupled together using hybrid coupling units are shown as follows:-

Figure 4	Shows the gain/frequency characteristic of the four individual amplifiers when driven to 300 Watts p.e.p. The variations of gain due to component and transistor spreads give a total gain variation between 14.5 and 18dB over the range f = 1.6MHz to f = 30MHz.
Figure 5	Shows the gain characteristic of the four amplifiers combined using hybrid couplers in the input and output circuits to give 1.0kW p.e.p. The gain variation is between 14dB and 17.5dB over the range f = 1.6MHz to f = 30MHz.
Figures 6 and 7	Show the total spread of intermodulation products (d_3 and d_5) over the range f = 1.6MHz to f = 30MHz of the four amplifiers driven to 300 Watts p.e.p. each. The intermodulation product level of each amplifier is better than -26dB for d_3 and better than -36dB for d_5.
Figures 8 and 9	Show the intermodulation product level of the complete system when driven to 1.0kW p.e.p. The lowest level for d_3 is -30dB and the lowest level for d_5 is -38dB over the entire range f = 1.6MHz to f = 30MHz.
Figures 10 and 11	Show the intermodulation product level of the complete system driven to 1.2kW p.e.p. The lowest level of d_3 intermodulation products is -26dB and of d_5 is -39dB.

3. THE HYBRID COUPLING CIRCUITS

3.1. Input Hybrid (Shown in Photograph 1)

3.1.1. The input hybrid coupler consists of four transmission line transformers wound on Mullard ferroxcube toroids type FX1588, each transformer using two stacked toroids.

3.1.2. The input transformers is 4:1 impedance from 50Ω unbalanced input to 12.5Ω unbalanced output and is wound with two parallel 50Ω cables (3mm ext diam) on the two stacked FX1588 toroids.

3.1.3. The second transformer is wound with nine turns of two parallel 50Ω cables (3mm ext. diam) with the connections arranged as a hybrid to give 12.5Ω to two 25Ω unbalanced in-phase outputs.

3.1.4. The third and fourth transformers are identical and are wound with 10 turns of 50Ω co-axial cable (3mm ext. diam) with the connections arranged as a hybrid to give 25Ω to two 50Ω unbalanced in-phase outputs.

3.1.5. Out of balance (or power dumping resistors) of 100Ω are connected across the two 50Ω output ports of each output transformer and a 50Ω resistor is connected across the two 25Ω output ports of the intermediate hybrid transformer.

3.2. Although theoretical transmission line transformers operate over a very wide band, there are, in practice, inevitable discontinuities caused by stray capacitance, stray leakage inductances etc. due to the construction of such a system. Also the relatively large ballast resistors and their associated connecting leads which have stray inductances must be compensated by capacitors to obtain a near constant impedance throughout the band f = 1.6MHz to f = 30MHz.

It was found experimentally, using a Hewlett Packard vector impedance meter that satisfactory impedance compensation could be obtained by the use of three capacitors (120pF, 80pF, 80pF) connected between the 12.5Ω terminal and earth and each of the two 25Ω impedance terminals to earth. The final result gave a maximum v.s.w.r. of 1.3:1 at the input port with all other ports terminated with 50Ω wideband loads. Figure 12 shows the block and circuit diagrams of the complete input hybrid; figure 13 shows measured results of v.s.w.r. throughout the band.

3.3. The rating of the hybrid resistors is sufficient for fail-safe operation of the system under the worst fault conditions for continuous c.w. operation without air cooling of the resistors.

4. THE OUTPUT HYBRID (Shown in Photograph 2)

4.1. The output hybrid consists of four transmission line transformers wound on Mullard ferroxcube toroids type FX1588. A pair of the amplifiers is coupled through 50Ω ports to a common 25Ω port. (Similarly for the other pair of amplifiers). Each of these transformers is wound with six turns of 50Ω P.T.F.E. dielectric co-axial cable (2.5mm ext. diam) on a stacked core of four FX1588 toroids, the connections arranged as a hybrid transformer from 50 plus 50 to 25Ω. Out-of-balance (or power dumping resistors) of 100Ω are connected across the two 50Ω ports of each hybrid transformer.

4.2. A third hybrid transformer couples the two 25Ω outputs to combine the power from these two sources to 12.5Ω impedance. This transformer is wound with four turns of two 50Ω P.T.F.E. dielectric co-axial cable (4.0mm ext. diam) on a stack of nine FX1588 toroids, the connections being arranged as a hybrid transformer from 25 plus 25Ω to 12.5Ω. An out-of-balance (or power dumping resistor) of 50Ω is connected across the two 25Ω inputs.

4.3. The fourth transformer is a 1:4 impedance transformer which transforms the combined output at 12.5Ω impedance to the 50Ω load impedance. This transformer is wound with five turns of two parallel 50Ω P.T.F.E. dielectric co-axial cables (4.0mm ext. diameter) on a stack of ten FX1588 toroids, the connections are arranged for a 1:4 impedance transformer (unbalanced).

4.4. To compensate the strays (leakage inductance and capacitance etc.) it was found experimentally, using a vector impedance meter, that satisfactory compensation could be obtained using five capacitors connected at the points shown in the circuit of Figure 14. The final result (Figure 15) shows a maximum v.s.w.r. of 1.16:1 on any input port with all other ports terminated with 50Ω wideband loads.

4.5. The rating of the hybrid resistors is sufficient for short term fault conditions with two-tone signals without forced air cooling. With suitable air cooling continuous c.w. operation at the most severe fault condition (two units inoperative) is possible.

5. PERFORMANCE OF THE 1kW AMPLIFIER

The complete system was assembled as in the block diagram of Figure 16.

The idling currents of the individual units were adjusted to 240mA total (2 x 120mA nominal) with the supply rail at 50 Volts.

The performance of the complete 1kW unit when driven by a high quality drive source (intermodulation products better than -45dB, and very low harmonic content) was as follows:

Two-tone signals driving to 500 Watt mean power

MHz	P.E.P. W	Ic_1 A	Ic_2 A	Ic_3 A	Ic_4 A	Ic_5 A	Ic_6 A	Ic_7 A	Ic_8 A	Ic_{tot} A	$P_{in\ tot}$ W	η_d %
1.6	1000	3.2	3.1	3.0	3.2	3.3	2.8	3.1	2.8	24.5	1223	40.1
3.5	1000	2.9	2.7	2.6	2.7	2.9	2.4	2.7	2.4	21.3	1060	47
5.0	1000	2.9	2.6	2.5	2.6	2.9	2.5	2.6	2.4	21.0	1050	47.5
7.0	1000	3.1	2.6	2.6	2.8	3.0	2.6	2.8	2.6	22.1	1108	45
10	1000	3.2	2.8	2.8	2.8	3.2	2.8	3.0	2.8	23.2	1160	43
14	1000	3.5	3.2	3.2	3.2	3.2	3.0	3.0	3.0	25.3	1260	40
18	1000	3.6	3.3	3.2	3.2	3.4	3.1	3.0	3.2	26.0	1300	38.5
20	1000	3.7	3.6	3.4	3.3	3.7	3.2	3.1	3.3	27.3	1360	36.5
24	1000	3.6	4.1	4.1	3.1	4.2	3.3	3.3	3.6	29.4	1470	34
28	1000	3.7	4.0	4.2	3.2	4.2	3.3	3.3	3.6	29.5	1475	33.9
30	1000	3.8	4.0	4.2	3.2	4.2	3.2	3.3	3.5	29.2	1460	34.2

6. NOTES ON THE MEASUREMENTS AND OPERATION OF THE 1kW AMPLIFIER

6.1. Power Measurements

The p.e.p. measurement has been derived from the mean power measurement using a 50dB wideband power attenuator and a Hewlett Packard calorimetric Wattmeter.

The p.e.p. was assumed to be twice the mean value; because the harmonic output varies with frequency and is particularly troublesome around 10MHz to 14MHz (see Figure 17) the error in this assumption introducted by harmonics could contribute approximately 25 Watts of mean power in the load at certain frequencies. Thus the true p.e.p. would then only be 950 Watts at the worst part of the band.

6.2. C.C.I.R. Recommendations

C.C.I.R. recommendations involve calorimetric c.w. power measurements at half power followed by the application of two-tone signals of the same amplitude as the c.w. signal. By observation of the two tone envelope on an oscilloscope the p.e.p. can be calculated from the relative amplitude of the observed waveform. Experience has shown that this method is also suspect owing to the distortion of the envelope waveform caused by the phase relationships of the harmonics. Discrepancies of 25% could result. The C.C.I.R. recommendations however, do allow the use of a filter, if necessary, with the oscilloscope. The use of such filters corrects the peak to mean relationship of the observed waveform but, of course, the harmonic power is still part of the mean load power.

6.3. Adjustment of Idling Current

The intermodulation distortion is affected by the idling current. The idling current was not critical for signals driving up to full p.e.p., but is somewhat critical for signals in the range -10dB to -20dB below maximum output. The idling current adjustment for optimum low level operation also varies with frequency and a resultant compromise was made at 240mA total for each amplifier.

6.4. Simulated Failure Operation

The amplifier has been run, and measurements made with as many as three of the amplifiers rendered inoperative. The hybrid coupling unit has performed satisfactorily with the exception that somewhat more power was measured in the load than is theoretically possible when one or two units were inoperative. This could be due to a reduced isolation between ports which was found, on measurement to deteriorate from -40dB at 3MHz to -20dB at 24MHz with the ports terminated by 50Ω. Inoperative amplifiers would not terminate the hybrids correctly and will therefore, cause a further decrease of isolation and hence cause the operating amplifier or amplifiers to deliver more power to the load and less in the dumping resistors.

7. OPERATING TEMPERATURE

The measurements on this amplifier have been made using a water cooled heatsink maintained at approximately 25°C. The ratings of the BLX15 allow operation up to 90°C heat sink temperature. However, at elevated temperatures, it has been established that there is an increase in intermodulation products at a given output level together with a reduction in the power gain. The following measurements were made on a single 300 Watt amplifier using two BLX15 transitors.

7.1. P_{load} 300 Watts P.E.P. I.P.'s v. Heatsink Temperature

f MHz	Heatsink temp. 25°C I.P.'s d3 dB	Heatsink temp. 50°C I.P.'s -d3 dB	Heatsink temp. 75°C I.P.'s d3 dB	Heatsink temp. 90°C I.P.'s d3 dB
1.6	-34	-34	-33	-33
5.0	-36	-35	-34	-33
10	-32	-31	-24	-28
20	-30	-29	-28	-27
30	-28	-27	-26	-25

7.2. P_{load} 300 Watts P.E.P. Power Gain v. Heatsink Temperature

f MHz	Gain at Heatsink temp. 25°C	Gain at Heatsink temp. 90°C
1.6	17.4dB	16.8dB
5.0	17.0dB	15.8dB
10	15.6dB	14.2dB
20	16.0dB	15.8dB
30	15.5dB	15.0dB

7.3. Output Power at Constant I.P. level v. Heatsink Temperature

f MHz	Heatsink temp. 25°C	Heatsink temp. 90°C
1.6	300	290
5.0	300	275
10	300	260
20	300	250
30	300	250

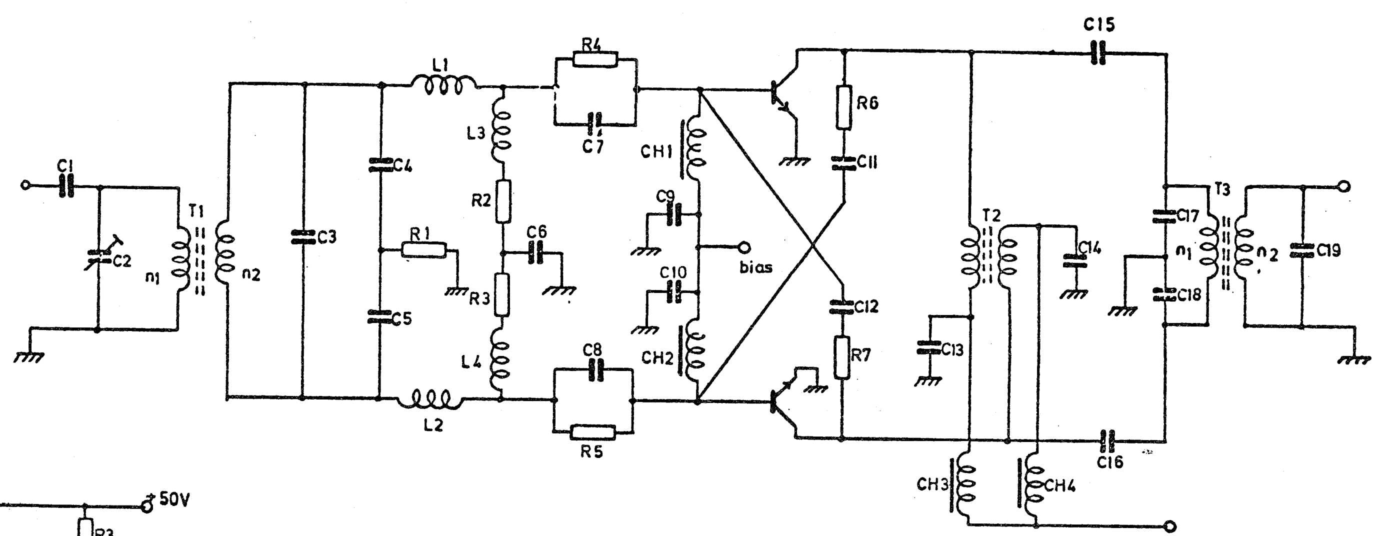
8. CONCLUSIONS

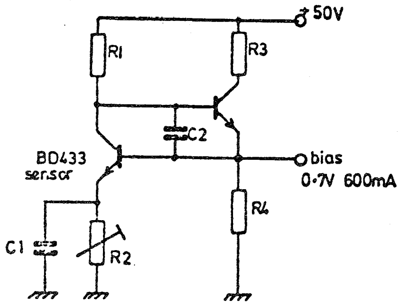
8.1. Four 300 Watt p.e.p. wideband amplifiers have been built and successfully combined using hybrid coupling units to provide 1kW p.e.p. from f = 1.6MHz to f = 30MHz with intermodulation products at -30dB at the worst part of the band.

8.2. Overdriving the amplifier to a power level of 1.2kW gives intermodulation products at -26dB at the worst frequencies.

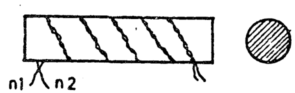
8.3. Although the output has been measured in a wideband load and therefore a small fraction of the measured power is provided by harmonics it would appear feasible to build a 1.0kW minimum output transmitter from four such 300 Watt BLX15 amplifiers.

8.4. The 1kW amplifier has been measured at a heatsink temperature of approximately 25°C. If operated at 90°C a deterioration of intermodulation products of 3dB must be expected. A power gain reduction of 1.5dB at mid-band would also occur. Alternatively a decrease of power output by 1dB for the same intermodulation products as at 25°C must be expected at 90°C.

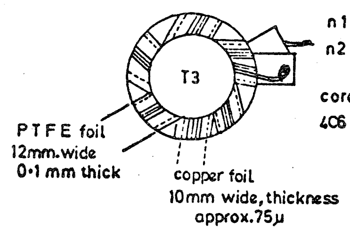




BIAS UNIT

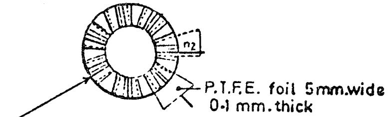


n1, n2, 5 turns of 2×1·0mm twisted enam.cu.wire
wound on a FXC rod grade 4B1 dimension 10×50mm.
part of antenna rod 3122 104 91250

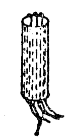


n1 = 6 turns, 2·07mm.en.cu.wire
n2 = 12 turns " " " " "

core: FXC toroid grade
4C6 36×23×15 3mm.
code 4322 020 91090



n1, 21 turns, 0·5mm en. cu. wire
n2, 7 turns, copper foil 3mm.wide app. 75μ thick
T1 FXC toroid grade 4C6 14×9×5mm
code 4322 020 91020



CL3 CL4
modified FXC choke grade 3B
code no. of bead 4132 020 31500

FIG.1.

Mullard Central Applications Laboratory Mitcham

COMPONENTS LIST

AMPLIFIER UNIT (Each Amplifier)

R_1	=	2 x 10Ω	in parallel	carbon	±5%	CR37
R_2 R_3	=	3 x 10Ω	in parallel	carbon	±5%	CR68
R_4 R_5	=	2 x 18Ω	in parallel	carbon	±5%	CR25
R_6 R_7	=	3 x 10Ω	in parallel	carbon	±5%	CR37

C_1	=	2 x 4.7nF	in parallel	polyester	±10%	347 series
C_2	=	100 pF [illegible] ceramic [illegible]				
$C_{\cancel{2}3}$	=	560pF		polystyrene	±1%	425 series
C_4 C_5	=	2 x 560pF	in parallel	polystyrene	±1%	425 series
C_6	=	100nF		polyester	±10%	347 series
C_7 C_8	=	3 x 1200pF		polystyrene		425 series
C_9 C_{10}	=	1µF moulded metallised polyester ±10% in parallel with 150µF tantalum 6V capacitor				347 series 421 series
*C_{11} C_{12}	=	82pF	ceramic, miniature plate		500V d.c.	
C_{13} C_{14}	=	3 x 100nF		polyester	±10%	347 series
C_{15} C_{16}	=	3 x 11nF		poly~~ester~~styrene	±1% 125V	436 series
*C_{17} C_{18}	=	2 x 180pF	in parallel	ceramic plate	500V d.c.	436 series
*C_{19}	=	2 x 22pF	in parallel	ceramic plate	500V d.c.	

T_1 See drawing, Figure 1
Philips toroid core grade 4 C6 Code No. 4322-020-91020

T_2 See drawing, Figure 1
Part of Philips antenna rod grade ~~4 B1~~ 4A10 Code No. ~~3122-104-91290~~ 4311 120 55390

T_3 See drawing on Figure 1
Philips toroid core grade 4 C6 Code No. 4322-020-91090

FIGURE 2

L_1 L_2	0.5 turn of 1mm enam. Cu wire Dint 6mm with 2 x 6mm leads
L_4 L_4	0.5 turn of 1mm enam. Cu wire Din 6mm with 2 x 7mm leads
*Ch_1 Ch_2	Ferroxcube choke grade 3B Philips Code No. 4312-020-36640
*Ch_3 Ch_4	Modified ferroxcube choke grade 3B (See drawing on Figure 1) Philips Code No. 4312-020-36640

*These components are not in Mullard range.

Bias Unit (See Circuit on Figure 1)

One bias unit is screwed to the side of each amplifier copper block and both transistors are isolated from the block by mica washers. A nylon screw and washer is used to isolate the BD433. One bias unit supplies the forward base bias for both transistors of each unit. The collector idling current is then adjusted by R_2.

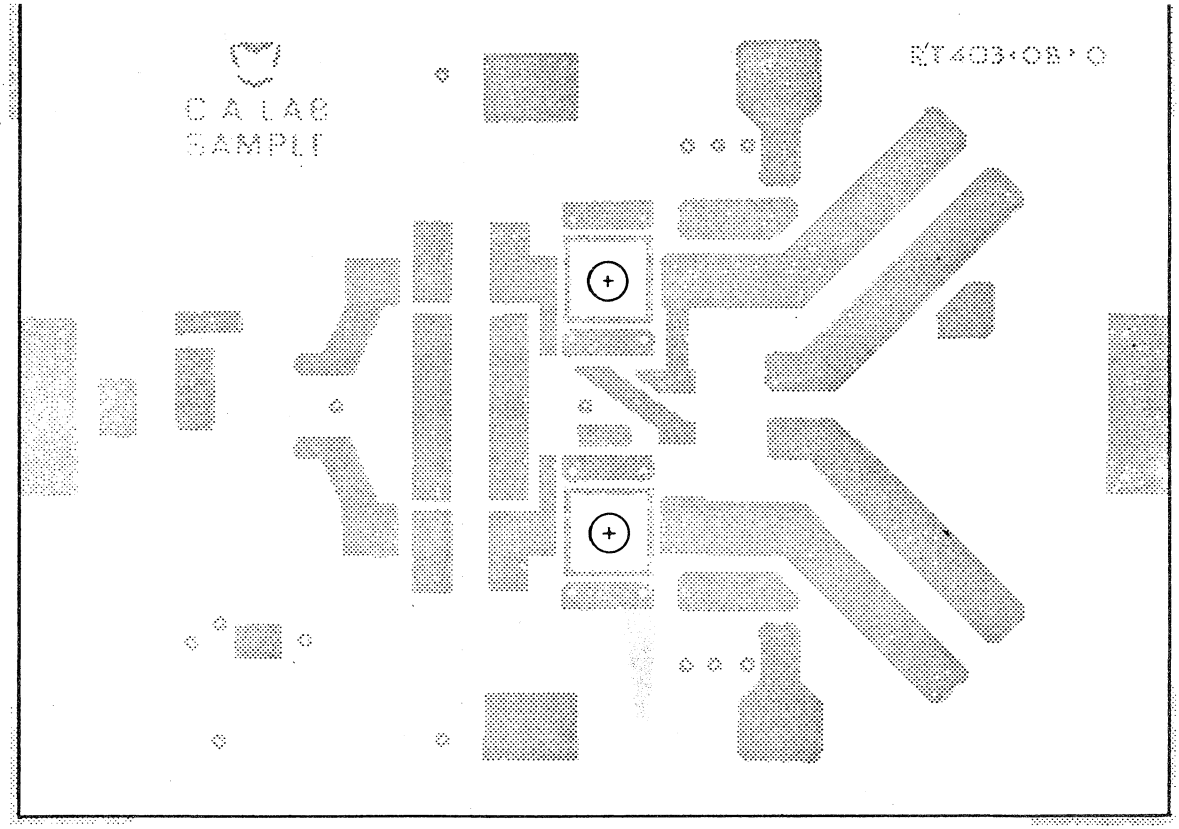
R_1	=	2.2kΩ	carbon	CR37
R_2	=	4.7Ω	3W pot w.w.	-
R_3	=	6 x 470Ω 5 Watt w.w. in parallel		
R_4	=	33Ω	carbon	CR37
C_1	=	100nF	polyester	347 series
C_2	=	100nF	polyester	347 series

Input Hybrid Coupler

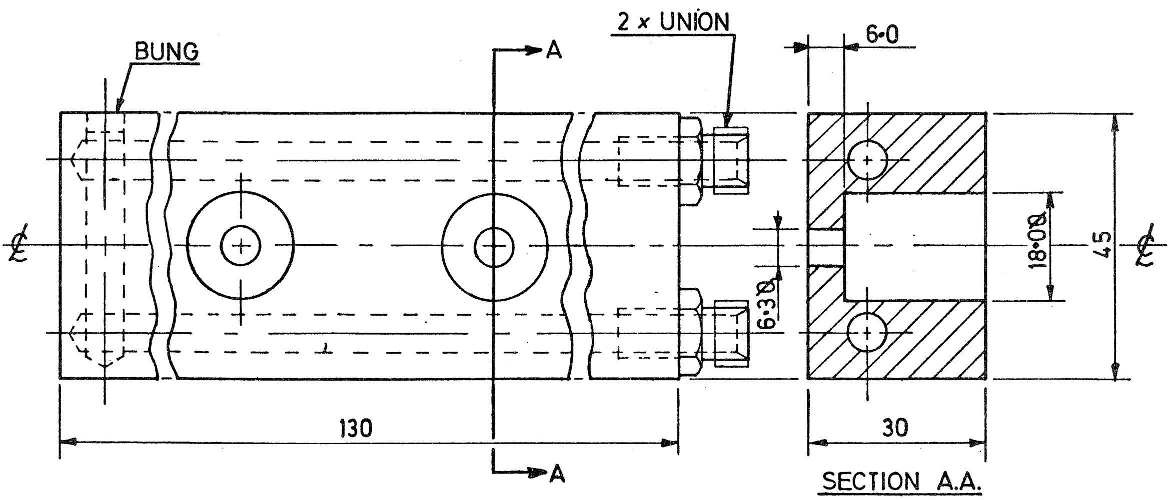
8	FX1588	Ferroxcube toroids	Mullard
1	50Ω	Electrosil resistor	Type H35 30W
2	100Ω	Electrosil resistor	Type H33 15W
	Miniature 50Ω co-axial cable windings		
	External diam. approx. 3mm from any suitable manufacturer		
2	82pF	tubular ceramic capacitors	low k
1	120pF	tubular ceramic capacitors	low k

Output Hybrid Coupler

27	FX1588	Ferroxcube toroids	Mullard
2	100Ω	Electrosil resistors in parallel	Type H37
2	100Ω	Electrosil resistors	Type H37
	50Ω P.T.F.E. dielectric insulated co-axial cable		
	approx. 4mm ext. diam. from any suitable manufacturer		
2	33pF	Ceramic block capacitors	A.T.C.
2	68pF	Ceramic block capacitors	A.T.C.
1	120pF	Ceramic block capacitors	A.T.C.



ACTUAL SIZE



COPPER BLOCK FOR WATER COOLING OR BOLTED TO ALUMINIUM FINNED EXTRUSION FOR FORCED AIR COOLING.

FIG. 3.

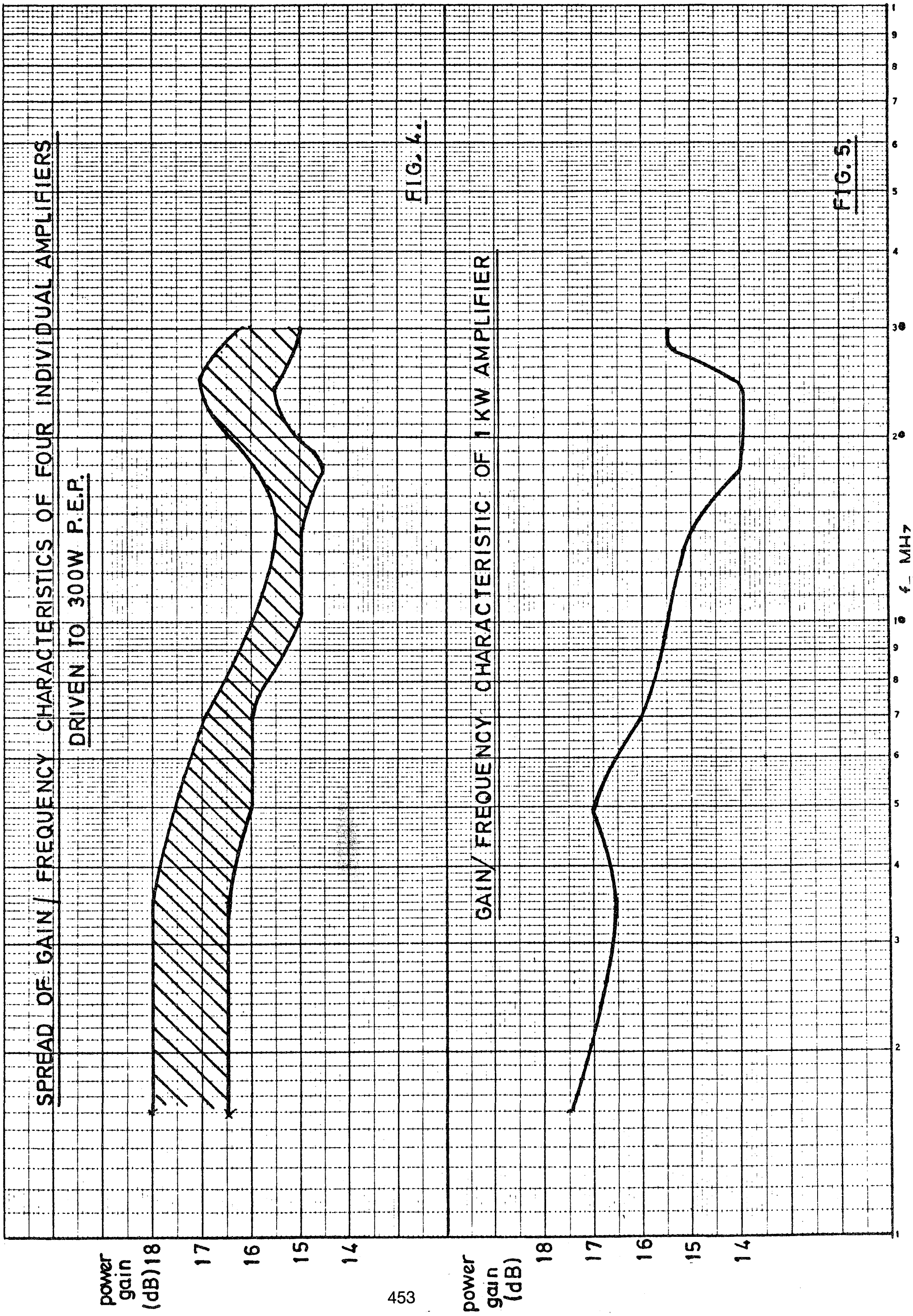
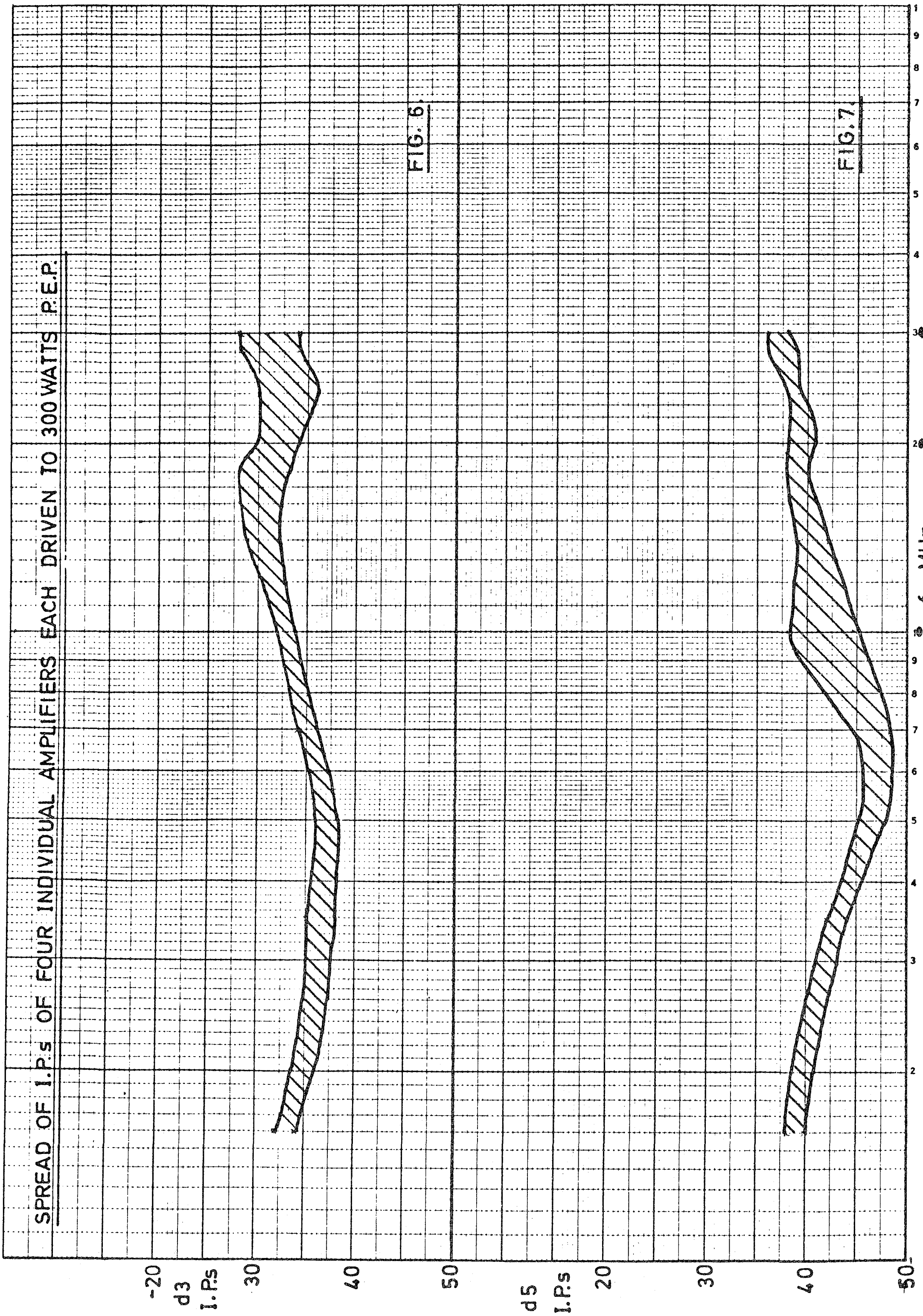
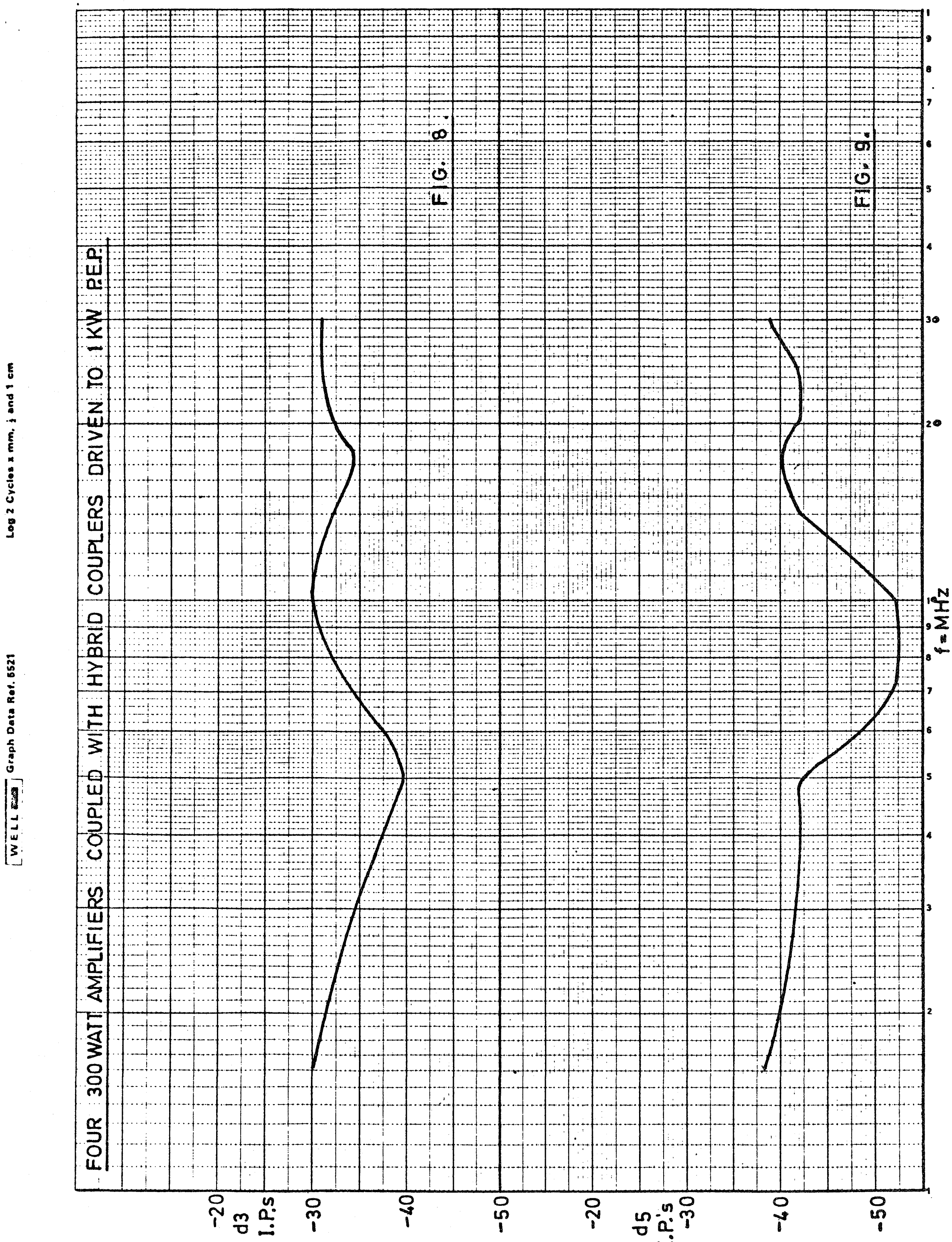

SPREAD OF GAIN/ FREQUENCY CHARACTERISTICS OF FOUR INDIVIDUAL AMPLIFIERS
DRIVEN TO 300W P.E.P.
GAIN/ FREQUENCY CHARACTERISTIC OF 1 KW AMPLIFIER
power gain (dB)
18
17
16
15
14
power gain (dB)
18
17
16
15
14
f_ MHz
1
2
3
4
5
6
7
8
9
10
20
30

FIG. 4.

FIG. 5.


SPREAD OF I.P.s OF FOUR INDIVIDUAL AMPLIFIERS EACH DRIVEN TO 300 WATTS P.E.P.
FIG. 6
FIG. 7
f = MHz
-20
d3
I.Ps
30
40
50
d5
I.Ps
20
30
40
50


FOUR 300 WATT AMPLIFIERS COUPLED WITH HYBRID COUPLERS DRIVEN TO 1 KW P.E.P.
FIG. 8.
FIG. 9.
f = MHz
d3 I.P.s
d5 I.P.'s
-20
-30
-40
-50
Graph Data Ref. 5521
Log 2 Cycles x mm, ½ and 1 cm

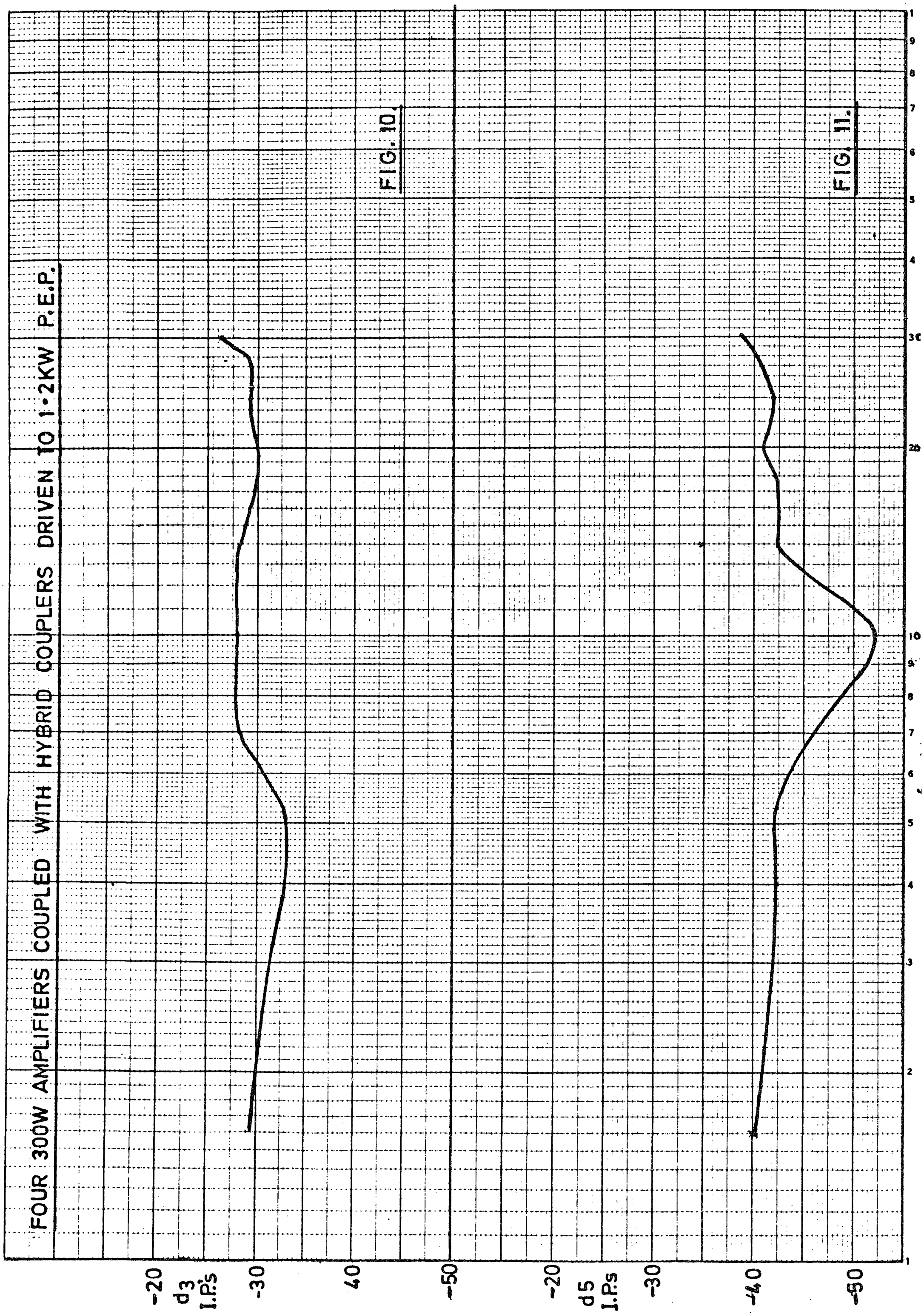
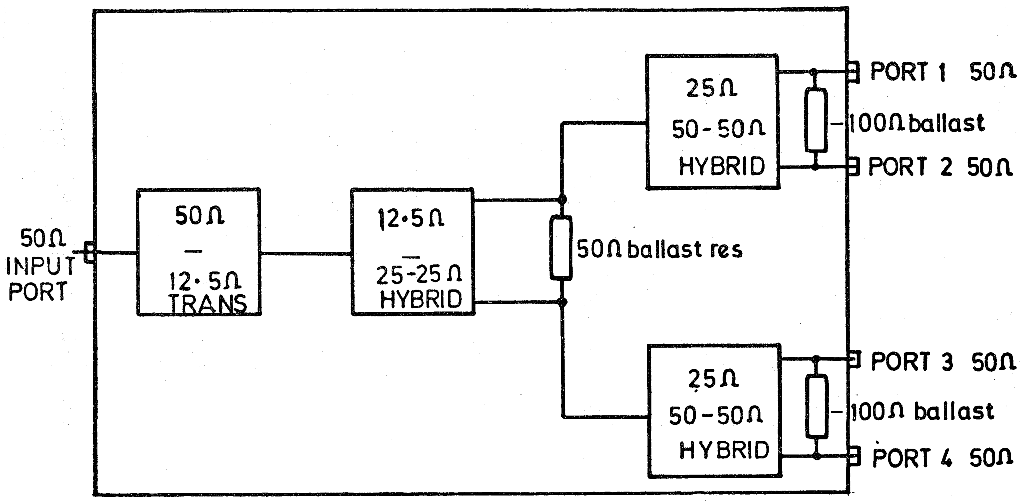

FOUR 300W AMPLIFIERS COUPLED WITH HYBRID COUPLERS DRIVEN TO 1·2KW P.E.P.
-20
d3
I.P.s
-30
40
-50
-20
d5
I.P.s
-30
-40
-50
1
2
3
4
5
6
7
8
9
10
20
30

FIG. 10.

FIG. 11.


50Ω
INPUT
PORT
50Ω
—
12·5Ω
TRANS
12•5Ω
—
25-25Ω
HYBRID
50Ω ballast res
25Ω
50-50Ω
HYBRID
PORT 1 50Ω
100Ω ballast
PORT 2 50Ω
25Ω
50-50Ω
HYBRID
PORT 3 50Ω
100Ω ballast
PORT 4 50Ω

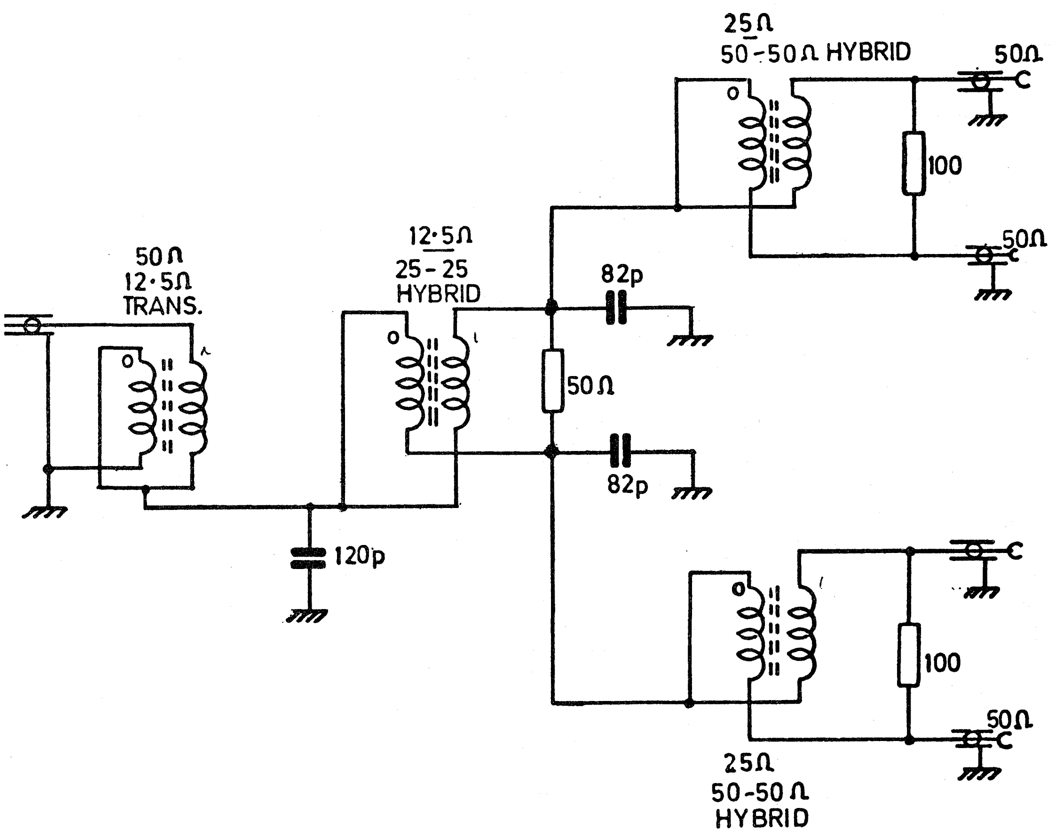
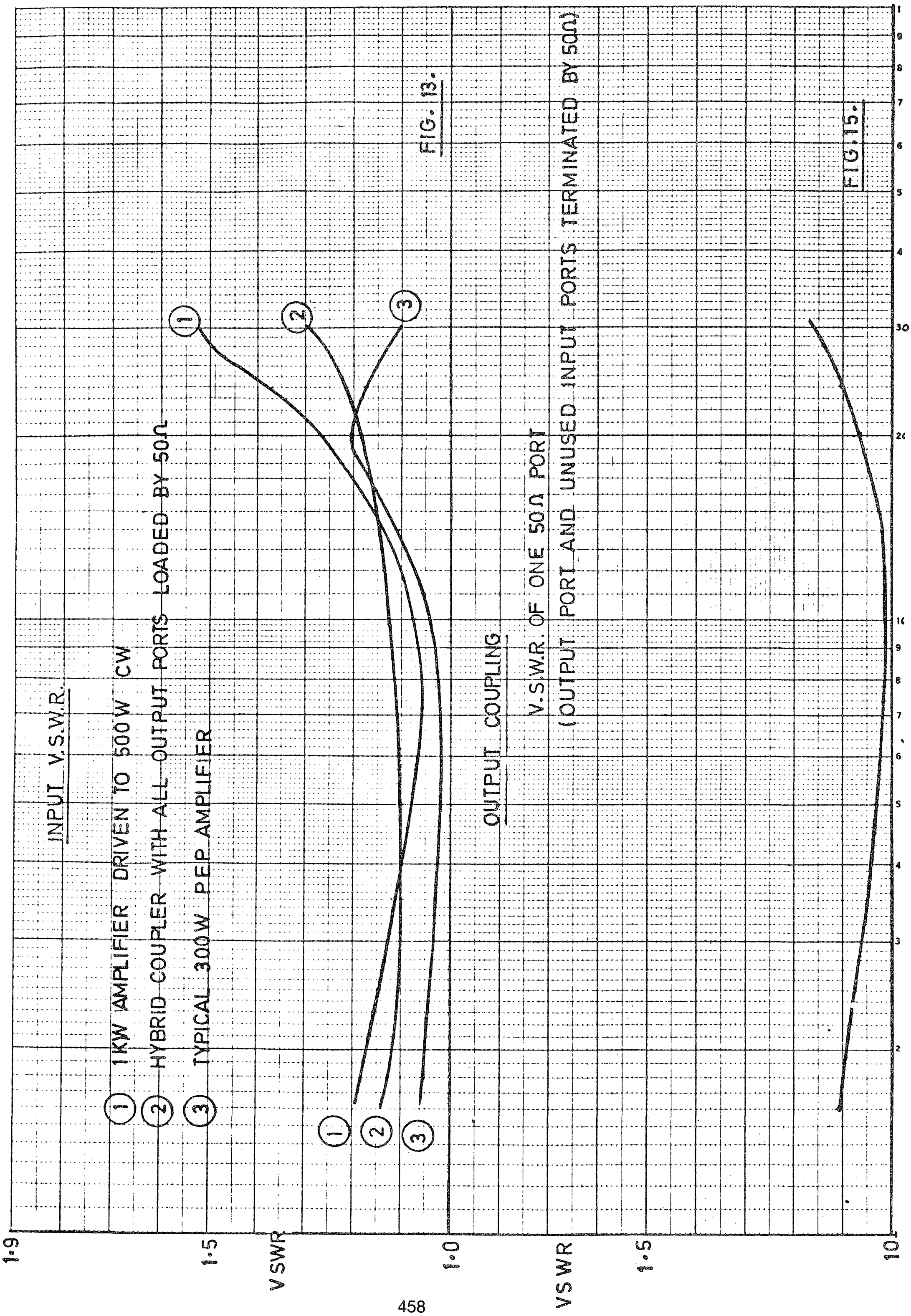

50Ω
12·5Ω
TRANS.
12·5Ω
25-25
HYBRID
120p
82p
50Ω
82p
25Ω
50-50Ω HYBRID
50Ω
100
50Ω
100
50Ω
25Ω
50-50Ω
HYBRID

FIG. 12.


INPUT V.S.W.R.
① 1KW AMPLIFIER DRIVEN TO 500W CW
② HYBRID COUPLER WITH ALL OUTPUT PORTS LOADED BY 50Ω
③ TYPICAL 300W PEP AMPLIFIER
FIG. 13.
OUTPUT COUPLING
V.S.W.R. OF ONE 50Ω PORT
(OUTPUT PORT AND UNUSED INPUT PORTS TERMINATED BY 50Ω)
FIG. 15.
1·9
1·5
VSWR
1·0
VSWR
1·5
1·0
1
2
3
4
5
6
7
8
9
10
20
30

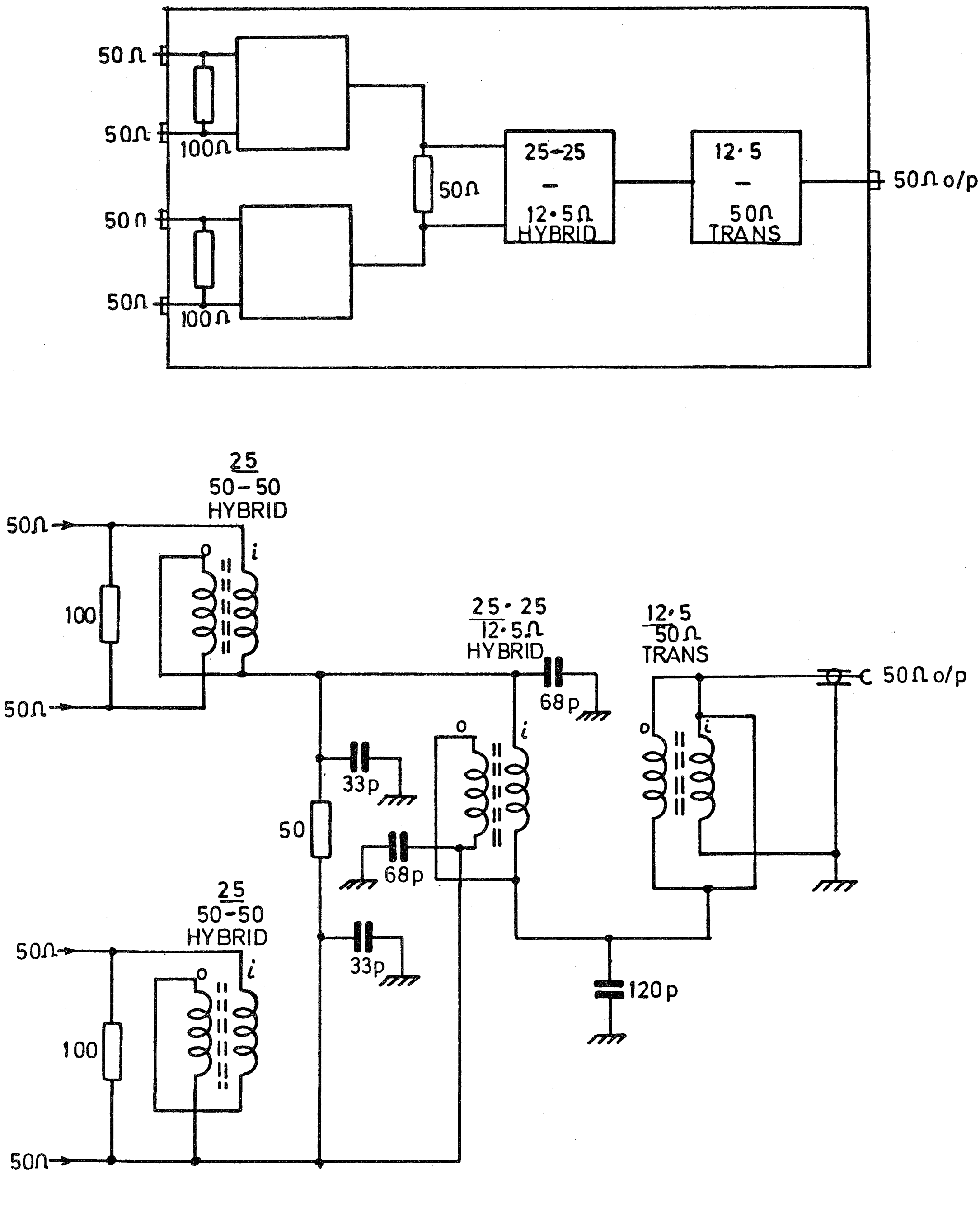

50Ω
50Ω
100Ω
50Ω
50Ω
100Ω
50Ω
25-25
—
12·5Ω
HYBRID
12·5
—
50Ω
TRANS
50Ω o/p
25
50-50
HYBRID
50Ω
100
50Ω
25·25
12·5Ω
HYBRID
68p
12·5
50Ω
TRANS
50Ω o/p
33p
50
68p
33p
120p
25
50-50
HYBRID
50Ω
100
50Ω

FIG. 14.

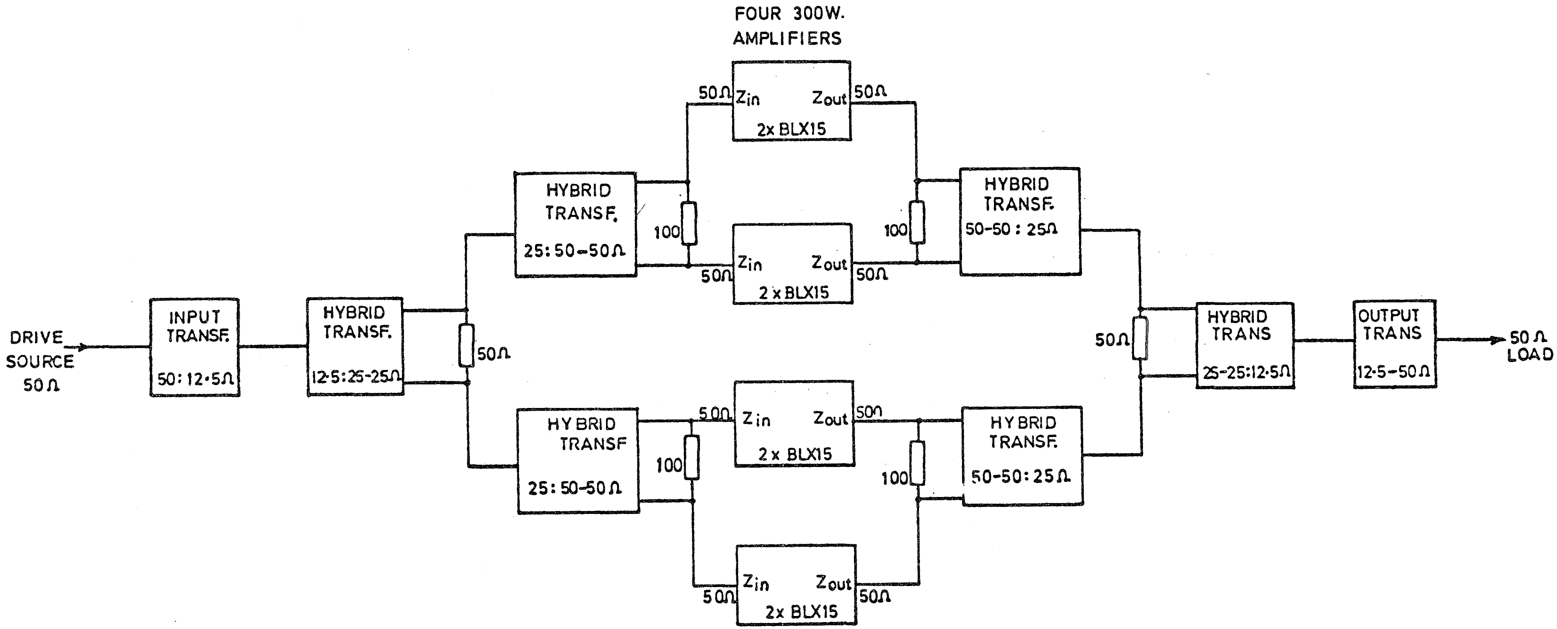
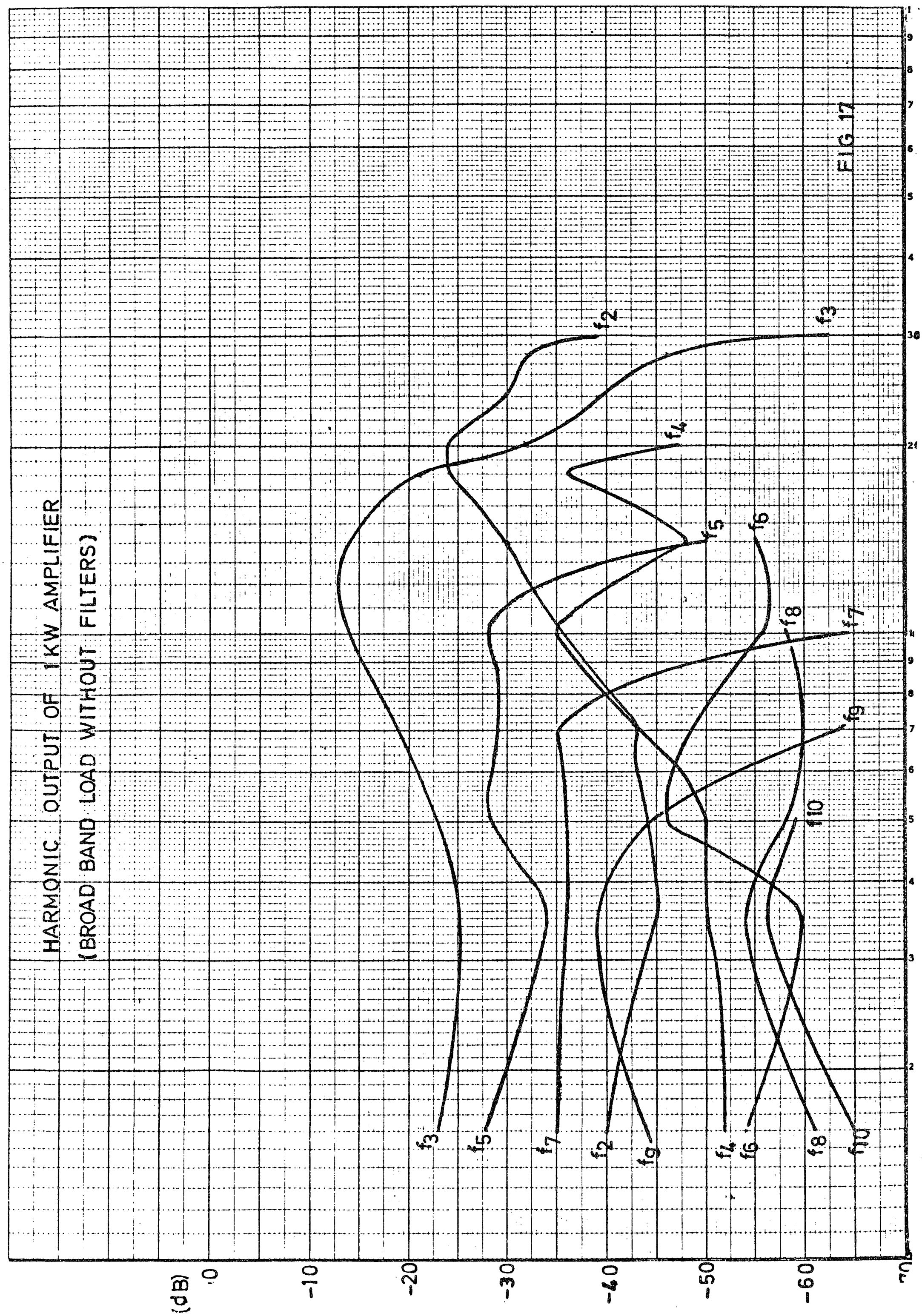
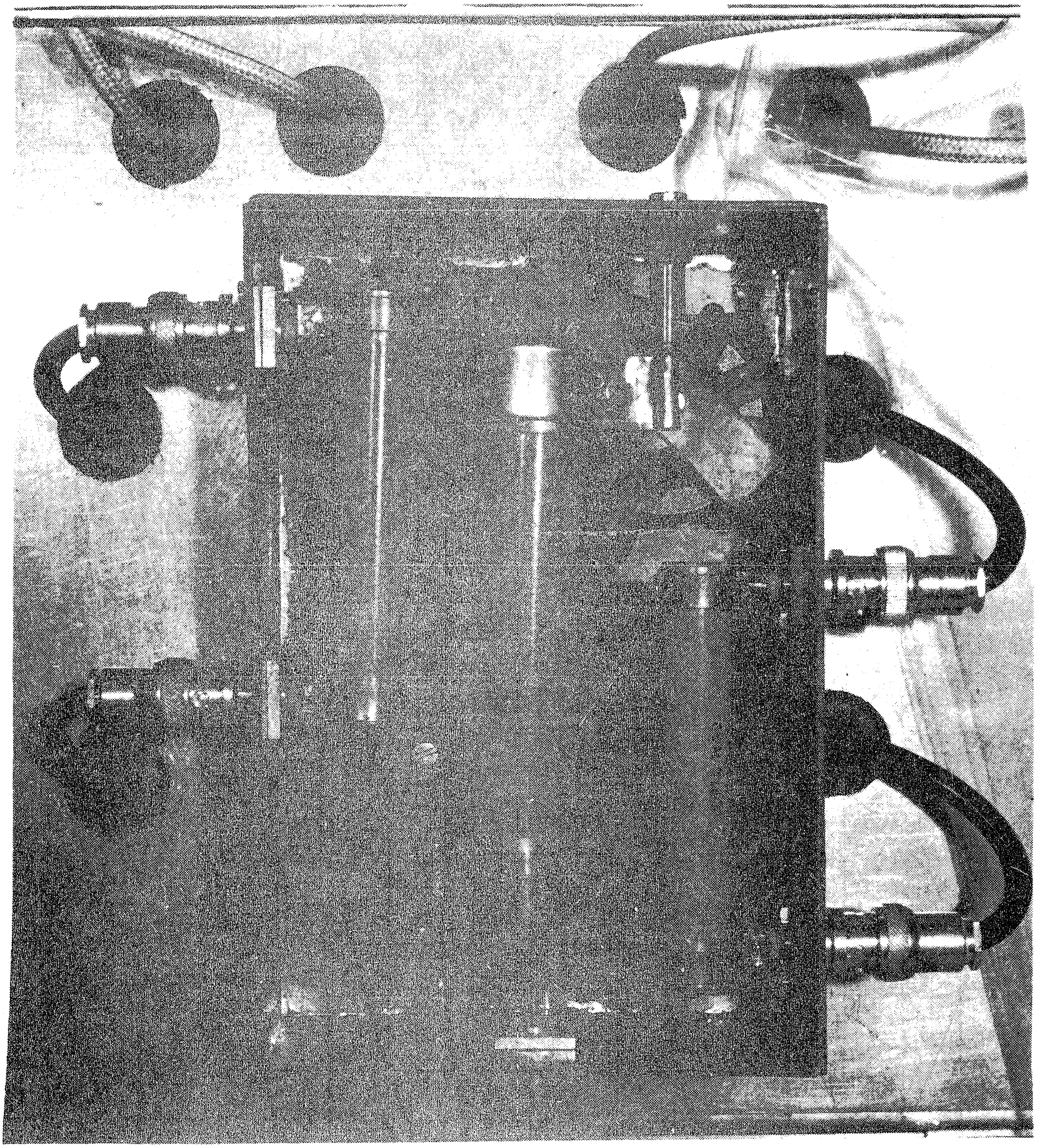


FIG. 16

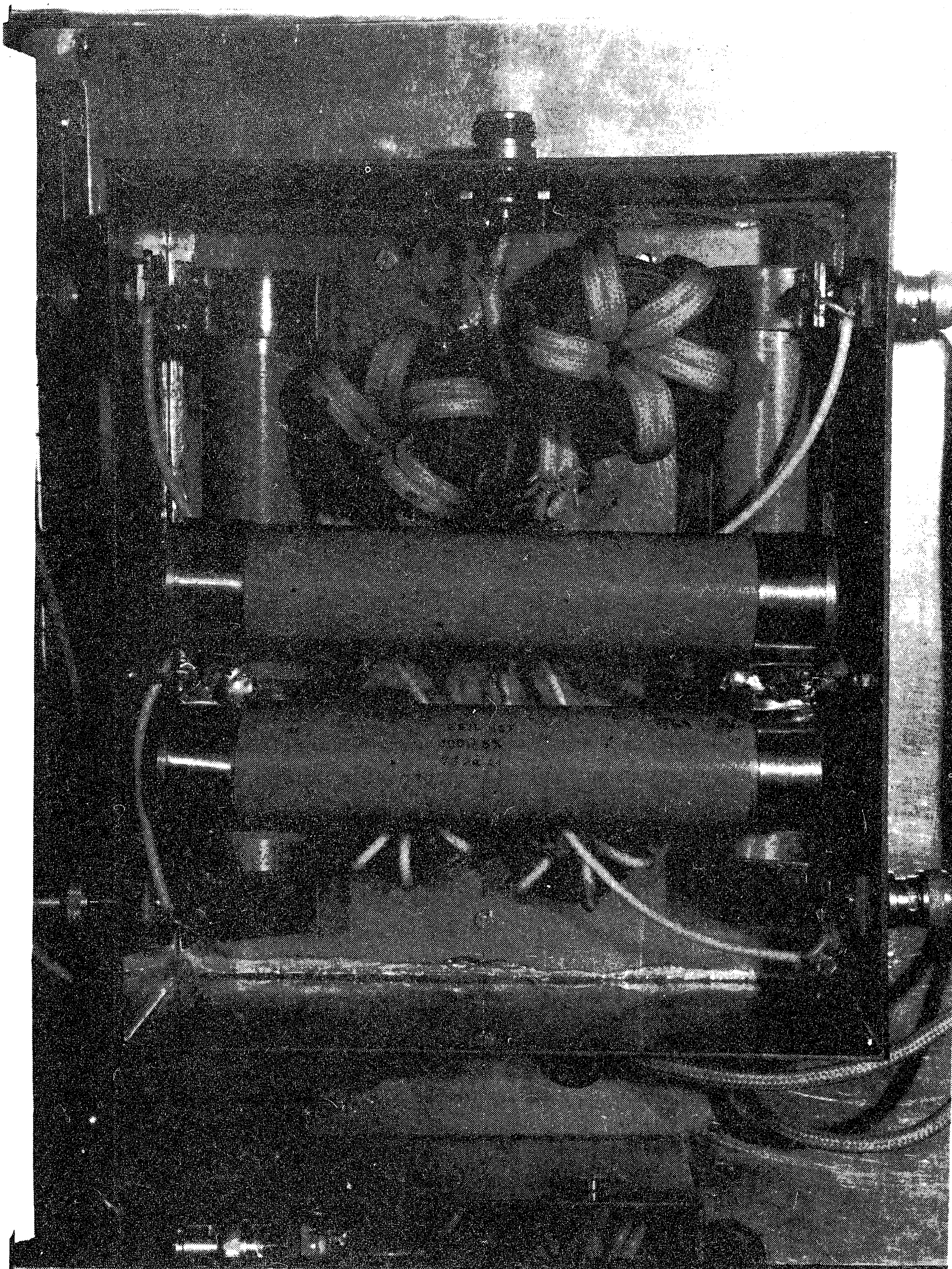
Mullard Central Applications Laboratory Mitcham


HARMONIC OUTPUT OF 1 KW AMPLIFIER
(BROAD BAND LOAD WITHOUT FILTERS)
FIG 17
(dB)
0
-10
-20
-30
-40
-50
-60
f2
f3
f4
f5
f6
f7
f8
f9
f10

THE INPUT HYBRID COUPLER



THE OUTPUT HYBRID COUPLER



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laboratory report

number : ECO7602 date : 6-4-1976

title : A High Quality Class A Linear H.F.Amplifier for Frequency Multiplex Operation using Cascoded BLW60 and ON495 Transistors.

author : A.Boekhoudt.

Electronic components and materials

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laboratory report

central application laboratory CAB
eindhoven - the netherlands

number: ECO 7602 | date: 6.4.1976
project: 6612 | pages: S 1 : A 4 : R 27

title: **A High Quality Class A Linear H.F. Amplifier For Frequency Multiplex Operation Using Cascoded BLW60 and ON495 Transistors**

author: A. Boekhoudt

summary: A description is given of a push-pull cascode class-A amplifier containing 2 pieces BLW60 and 2 pieces ON495. It is able to produce an output power of 50W P.E.P. in the frequency band of 1.6 to 28MHz.
All intermodulation and harmonic products are below -40dB. The amplifier has a power gain of 18.07 ± 0.23dB, whilst the input V.S.W.R. is 1.24 max. The required supply voltage is 44V and the current consumption appr. 6A.

A.H. Hilbers

Advies Octrooi d.d. 21 apr. 1976	X AV	GV		B		BL
Opgave Mamo d.d. 21 apr. 1976	X AV	X GV	X SP	B		BL
Datum: 07 apr. 1976	Mamo					

Abstract:

For S.S.B. transmitters in the frequency range of 1.6 to 28MHz frequency multiplex operation is a desirable feature. This implies that the intermodulation distortion and harmonic level of the power amplifier should be -40dB or better.

The push-pull cascode amplifier described in this report meets this target for an output power of 50W P.E.P. It contains 2 transistors type BLW60 and 2 transistors type ON495 (585BLY).
The low harmonic level in the upper part of the frequency band has been obtained with an additional low-pass filter (f_{co} = 29MHz) at the output of the amplifier.

Parasitic oscillation of the common-base section has effectively been suppressed by application of a high-pass filter section between the collectors of the ON495's.

At the input of the amplifier a correction network has been used that reduces the input V.S.W.R. to below 1.24. In addition it equalizes the power gain to 18.07 ± 0.23dB.

The required supply voltage is 44V and the current consumption appr. 6A.

The circuit diagram is depicted in fig. A, whilst fig. B shows the power gain, input V.S.W.R. and I.M. distortion versus frequency. Suppression of the 2nd and 3rd harmonic versus frequency can be found in Fig. C.

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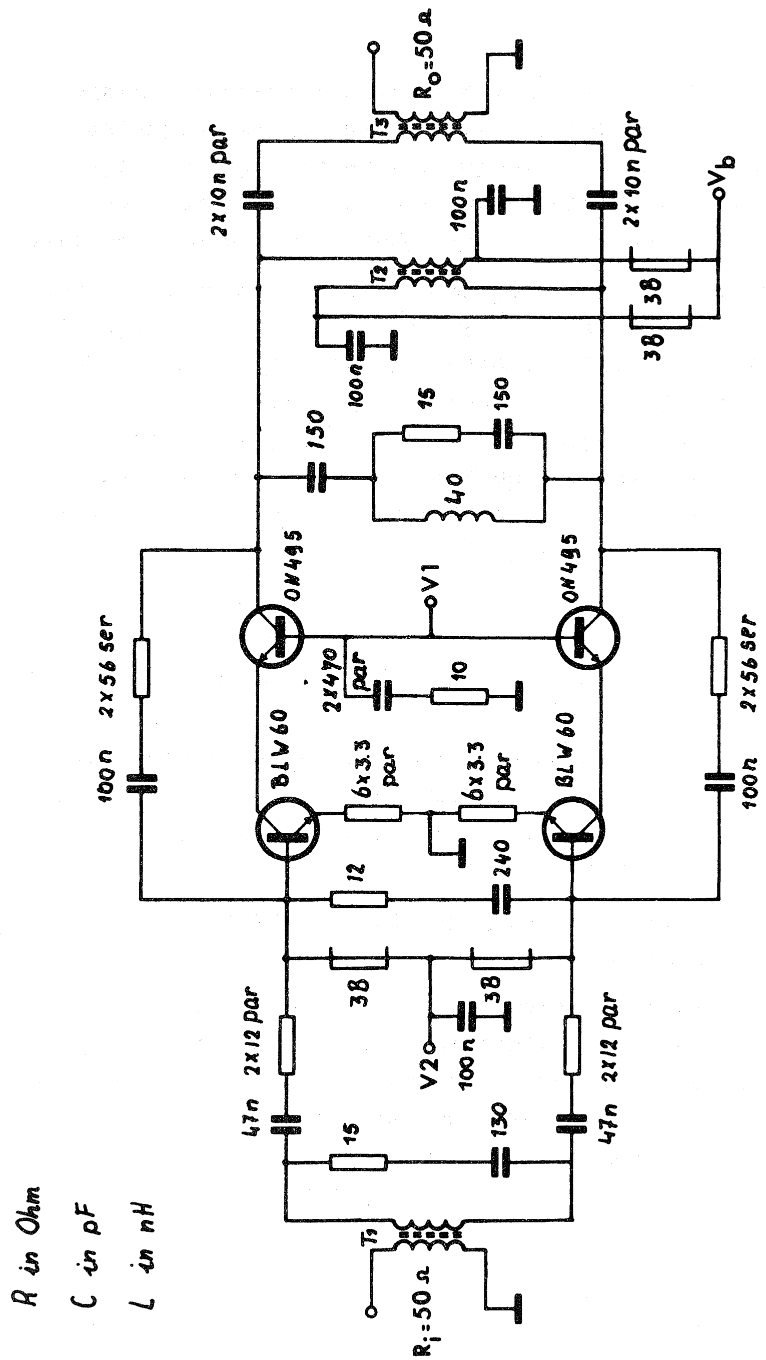

Figure A

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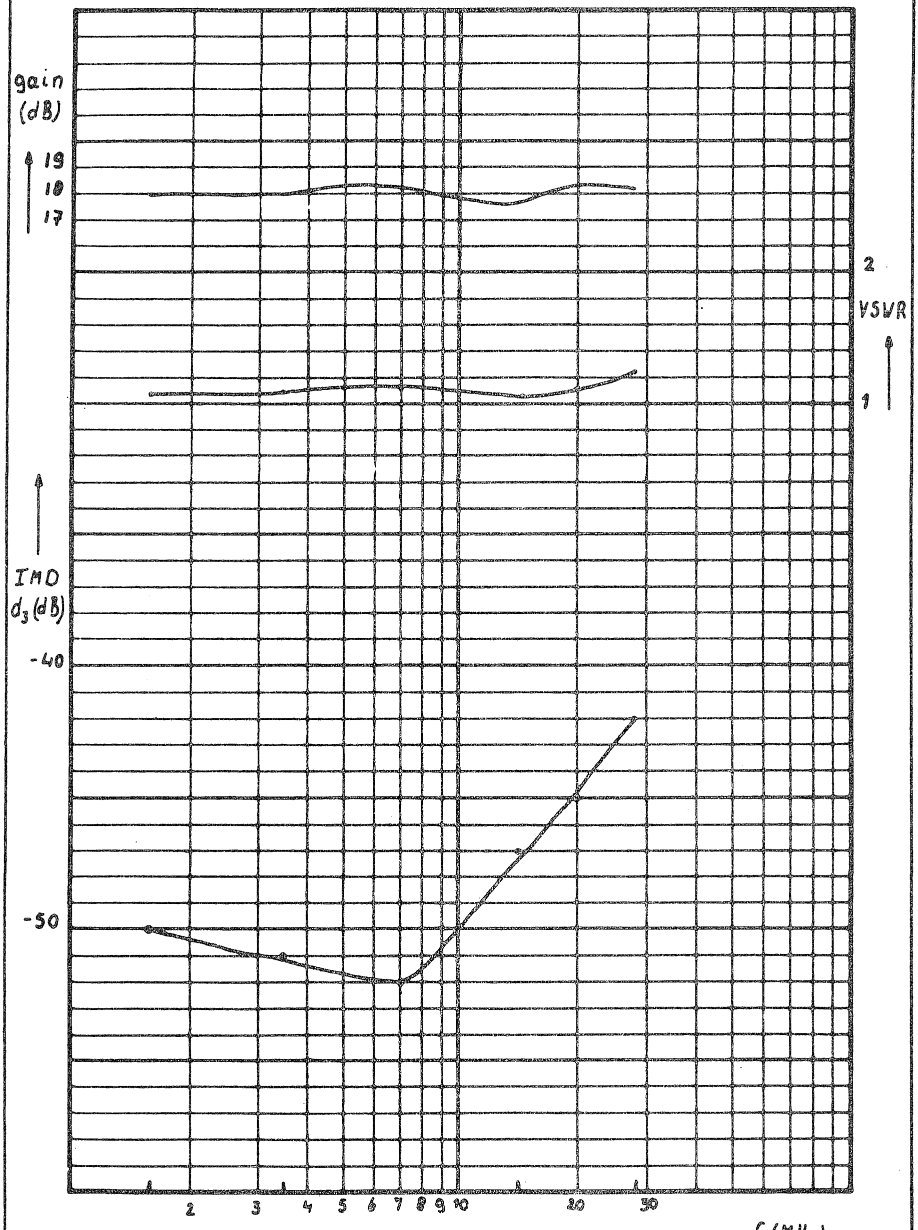

figure B

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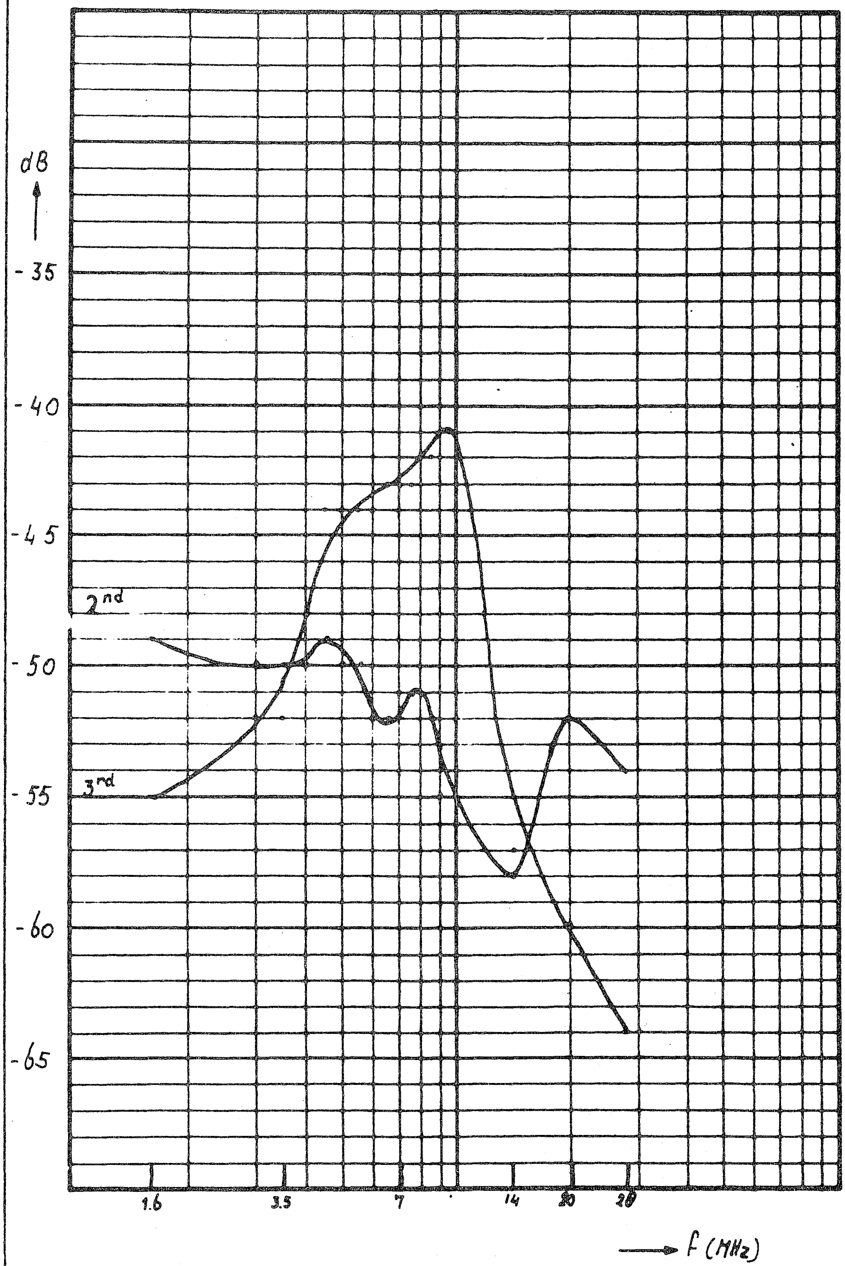

Figure C

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1. Introduction

To apply multi-channel communication with S.S.B. transmitters it is required to have a wideband power amplifier of which all intermodulation and harmonic distortion products are suppressed by at least 40dB.

Existing class-AB amplifiers typically have an I.M. distortion of -30dB and a third harmonic suppression of only 15dB. Moving to class-A operation will reduce the I.M. distortion to -40dB, however the harmonic suppression is still not better than appr. 25dB.

This report describes a broadband cascode push-pull amplifier that meets the above mentioned requirements. The cascode consists of 2 grounded emitter BLW60's driving 2 grounded base ON495's (585BLY) which deliver 50W P.E.P. with a gain of 18dB over the band 1.6 to 28MHz. The supply voltage is 44 Volts.

2. Design considerations

The main causes of distortion in linear amplifiers are:

a. variations of f_T versus collector current and voltage and

b. variation of collector capacitance versus collector voltage.

In class-A common emitter amplifiers point b. gives the highest contribution to non-linearity in the upper part of the frequency band. This is because the R.F. current through the collector capacitance is the larger part of the input current.

In a cascode circuit the situation is quite different as the common emitter stage is loaded with the very low input impedance of the common base stage. This causes a reduction of the current through the collector capacitance by appr. a factor 10.

In the common base section the influence of the collector capacitance is also small, but for another reason. Here it is in parallel with the load where its reactance is high compared with the load resistance.

These are the reasons why the cascode circuit has been succesfully applied in CATV-amplifiers.

The transistors are connected in series for D.C., with a V_{CE} of 14V for the BLW60, 28V for the ON495 and about 2V for the emitter resistor of the BLW60, so a D.C. supply voltage of about 44V is needed. The D.C. SOAR of both BLW60 and ON495 allow a collector current of 3A at a heatsink temperature of 70°C.

The first set-up was a single-ended amplifier with two transistors in cascode, to get an idea of the possiblities with this type of circuit.

The target for this amplifier was an output power of 20W P.E.P., with harmonic and intermodulation levels below -40dB. The experiences obtained with this single amplifier were used in the cascode push-pull amplifier.

To get a better harmonic suppression in the upper part of the band a Chebyshev low-pass filter with a cut-off frequency of 29MHz has been applied at the output of the amplifier.

3. Circuit description

3.1 Basic circuit

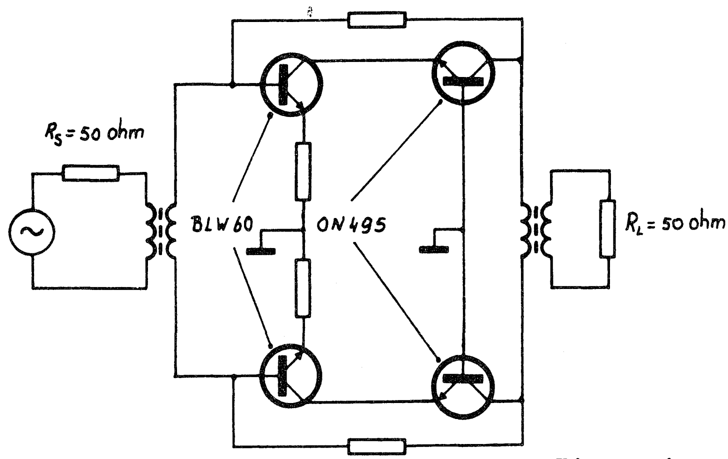

Figure 1

With the aid of the formules and rules mentioned in C.A.B. report ECO 7201 [1] the currents, voltages and components of the cascode push-pull amplifier can be calculated.
Fig. 2 shows a simplified H.F. circuit of the single amplifier.

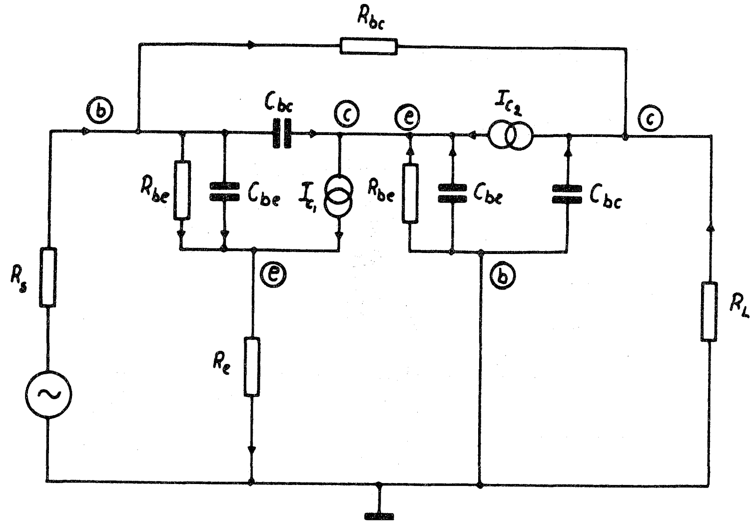

Figure 2

$I_{c_{be}}$ can be calculated from the relation

$f_T/f = I_{c_1}/I_{c_{be}}$

As the h_{FE} of the ON495 is 50, I_{c_1} = 3,06A.
Further the f_T of the BLW60 is 600MHz, so:
$I_{c_{be}} = \frac{28}{600} \cdot 3{,}06 = 0{,}143$A at f = 28MHz.

C_{bc} of the BLW60 is 83pF at V_c = 14V. The A.C. voltage across C_{bc} consists of: the voltage across R_e (2V), the V_{be}'s of both transistors (0,06V) and the voltage drop across R_c of the BLW60 plus R_e of the ON495 (0,6V).

In addition to this there is some voltage drop across the parasitic inductances in the circuit, so the total A.C. voltage has been estimated at 3V.

By this voltage excitation the effective capacitance of C_{bc} will increase by appr. 5%, so it becomes 88pF. Now the R.F. current at 28MHz can be calculated:

$I_{C_{bc}} = V_{bc} \omega C_{bc} = 3 . 2\pi\ 28.10^{6} . 88.10^{-12} = 0{,}046A.$

The total "imaginary" input current is:

$I_{C_{be}} + I_{C_{bc}} = 0{,}143 + 0{,}046 = 0{,}189A.$

The total "real" current must be at least 1.5 times this figure to obtain good wideband properties.

The input voltage is equal to the voltage drop across R_e (2V) plus the V_{be} of the BLW60 (0,03V), so $V_i = 2{,}03V$.

If we choose an input resistance of 6.25 ohm, the input current becomes:

$I_i = 2{,}03/6{,}25 = 0{,}325A.$

which is sufficient from the bandwidth point of view. Part of this current is consumed by R_{BE}:

$I_{R_{be}} = I_{c1}/h_{FE} = 3{,}06/50 = 0{,}061A,$

so the current through R_{bc} must be:

$I_{R_{bc}} = 0{,}325 - 0{,}061 = 0{,}264A.$

The voltage across this resistor is:

$V_{R_{bc}} = V_o + V_i = 28 + 2{,}03 \simeq 30V,$

so $R_{bc} = 30/0{,}264 = 114$ ohms.

Now R_e can also be calculated. As $V_{R_e} = 2V$ and the emitter current of the BLW60 is

$3{,}06 + 0{,}06 = 3{,}12A$: $R_e = 2/3{,}12 = 0{,}641$ ohm.

The internal R_e of the BLW60 is 0,088 ohm, so the external resistance must be 0,641 - 0,088 = 0,553 ohm.

The R.F. current delivered to the load is:

$I_o = I_{c2} - I_{R_{bc}} = 3 - 0{,}264 = 2{,}74A.$

From this the maximum output power and optimum load resistance can be calculated:

$$P_o = \frac{V_{c2} \cdot I_o}{2} = \frac{28 \cdot 2{,}74}{2} = 38{,}4W \text{ and}$$

$$R_L = V_{c2}/I_o = 28/2{,}74 = 10.2 \text{ ohms.}$$

The drive power is:

$$P_{dr} = \frac{V_i \cdot I_i}{2} = \frac{2{,}03 \cdot 0{,}325}{2} = 0{,}33W.$$

So the power gain is:

$$G_p = 10 \log P_o/P_{dr} = 10 \log \frac{38{,}4}{0{,}33} = 20.7dB.$$

3.2 The input transformer

Based on the calculations of the previous section a transformer should be needed for matching a 50 ohms source to 2x6,25 ohms. As a first experiment this has been done. However both power gain and input impedance were higher than expected in the upper part of the frequency range. To make corrections possible resistors of 6 ohms have been connected in series with the bases of the BLW60's and so the matching has been made from 50 ohms to 2x12.25 ohms. These corrections will be discussed in a later section. The input transformer is of the conventional type.
Because a good explanation of calculations on transformers is given in the C.A.B. reports ECO 6907 [2] and ECO 7213 [3], a description has been omitted in this report.

3.3 The output transformer

In this case a load impedance of 2x10.2 ohms has to be matched to 50 ohms. This transformer is also of the conventional type and the method of calculation can again be found in Ref. 2 and 3.

3.4 The centre tapped choke

For a description one is referred to C.A.B. report ECO 7114 [4].

In this amplifier the centre tapped choke is adapted to the optimum load resistance of appr. 10 ohm per transistor.

3.5 Compensation of the output circuit

The total collector load of the ON495's is formed by the circuit shown in Figure 3 in which C_o is the transistor output capacitance and C a d.c. blocking capacitor. The value of $L_{CH} = L_{TR}$ amounts to appr. 8 μH.

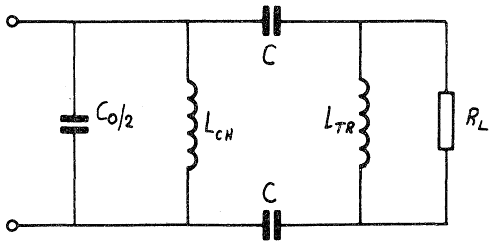

Figure 3

$$\text{So } C = \frac{L}{R^2} = \frac{8.10^{-6}}{20^2} = 20\text{nF}.$$

3.6 The bias circuit

The bases of the BLW60's and the ON495's are biased with two small stabilizers (see Figure 16). The base of the BLW60 needs a voltage of appr. 2.7 Volt ($= V_2$) because the voltage drop across R_e is 2 Volt and $V_{be1} \simeq 0.7V$.
The base of the ON495 is biased at 16.7V ($=V_1$): $V_{be2} \simeq 0.7V$ and $V_{e2} = V_{ce1} + V_{Re1} = 14 + 2 = 16V$.
The current to the base of each transistor is

$\frac{I_c}{h_{FE}} = \frac{3}{50} = 60mA$ typ., so the total current per supply is typ. 120mA.

3.7 The high-pass filter

For a better harmonic suppression a low-pass Chebyhev filter is used at the output. This filter has the disadvantage of increasing the load impedance outside the band, by which the amplifier starts oscillating.
A solution was found by connecting a simple high-pass filter between the collectors of the ON495's, which takes over the load from the Chebyshev filter above about 40MHz. See also Figure 4.
The capacitor C_1 of this network also functions as a primary H.F. compensation of the output transformer.
The secondary compensation capacitor has been left out because of better I.M. results at 28MHz.
Due to the presence of the high-pass section, the low-pass filter has to be of the L-input type.

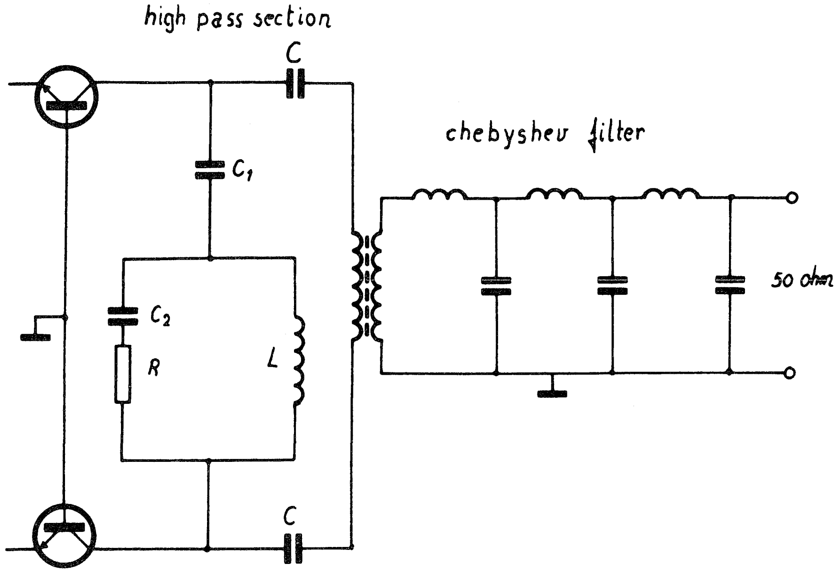

Figure 4

3.8 Correction of the input section

The first measurements on the push-pull amplifier without resistors in series with the bases of the BLW60's showed higher power gain and input impedance in the upper part of the frequency band than expected. In addition the input impedance was inductive rather than capacitive. As this impedance can be approximated by:

$$Z_i = \frac{Z_E \cdot R_{BC}}{Z_L}$$

its behaviour can be understood because:

a. the emitter resistors have series inductance and

b. the collector load impedance has a capacitive component (high-pass section) which can not be neglected.

The solution to this problem was found by:

a. application of resistors in series with the bases of the BLW60's.

b. modifying the input transformer accordingly and

c. adding 2 RC series combinations, one across the secondary of the input transformer and the other between the bases of the BLW60's.

Without further H.F. compensation on the primary side of the input transformer a sufficiently flat power gain versus frequency combined with a low input V.S.W.R. were obtained in this way.

4. The practical amplifier

Figure 9 shows the circuit diagram of the amplifier. The components were mounted on a double clad P.C. board of epoxy-glass with dimensions $215 \times 120 \times 1.6 mm^3$.

The interconnections and construction can be seen in Figures 17, 18 and 19 where the lower sheet functions as a ground plane. Connections of some upper parts with the ground plane were made with 2mm tubular rivets being soldered to the tracks to be sure of contact.

The transistors were water-cooled and provided with thermally conducting paste.

To avoid that H.F. power will penetrate into the bias supplies both are decoupled. The one for the BLW60's with 100nF's, the other for the ON495's with an RC combination of 10 ohm in series with the parllel connection of 2x 470pF.

The components of the bias network are mounted on a single clad PC board of epoxy-glass with dimensions 80x40x1.6mm³ (see Figures 11 and 15).

5. Measurements

5.1 General

The measurements were done under nominal conditions which implies:
Battery voltage V_B = 44 Volt
Load and source impedance R_L, R_S = 50 ohm
Ambient temperature T_{amb} = 25°C.
A single and a push-pull amplifier have been made for respectively 20W and 50W output power.
From both the measuring results will be given.

5.2 Intermodulation distortion

The I.M.D. versus frequency is measured with a two-tone signal (p, q). For the test set-up see Figure 5.
Table 1a+b presents the measuring results of the single and the push-pull amplifier.

Results of the single amplifier with P_o = 20W P.E.P.

f(MHz)	d_5	d_3	d_3	d_5(-dB)
1.6	- *)	53	53	-
3.5	-	-	-	-
7	-	55	-	-
14	-	49	54	-
20	-	45	45	-
28	57	42	41	56

Table 1a

Results of the push-pull version with P_o = 50W P.E.P.

f(MHz)	d_5	d_3	d_3	d_5 (-dB)
1.6	-	50	50	-
3.5	-	51	52	-
7	-	52	53	-
14	-	47	47	-
20	-	45	45	-
28	55	42	42	55

Table 1b

*) Means less than -60dB.

Fig. 7 shows d_3 (worst component) versus frequency for the push-pull version.

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Double tone test set up

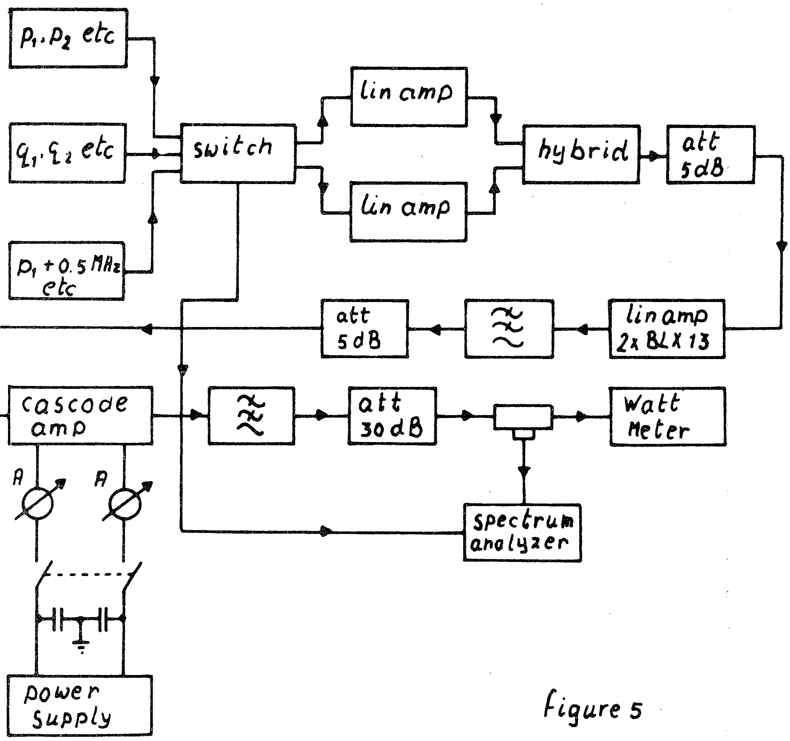

figure 5

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5.3 Single-tone measurements

With a single tone the P_i, power gain, input V.S.W.R. and harmonic content have been measured. Table 2a gives the results of the single amplifier for P_o = 20W, whilst table 2b gives the results of the push-pull amplifier for P_o = 50W.

f(MHz)	P_i (W)	Gain (dB)	VSWR	Harmonic 2nd	Harmonic 3rd (-dB)
1.6	0.41	16.88	1.12	51	59
3.5	0.41	16.88	1.09	51	52
7	0.41	16.88	1.13	43	50
14	0.41	16.88	1.20	45	62
20	0.40	16.99	1.14	55	70
28	0.47	16.29	1.17	56	- **)

Table 2a

f(MHz)	P_i (W)	Gain (dB)	VSWR	Harmonic 2nd	Harmonic 3rd (-dB)
1.6	0.79	18.01	1.08	49	55
3.5	0.78	18.07	1.08	50	51
7	0.77	18.12	1.13	51	43
14	0.82	17.85	1.05	58	57
20	0.74	18.30	1.12	52	60
28	0.77	18.12	1.24	54	64

Table 2b

**) Means less than -70dB

The test set-up is given in Figure 6, whilst Figure 7 shows the V.S.W.R. and gain versus frequency for the push-pull amplifier. Figure 8 gives the harmonic content versus frequency also for the push-pull version.

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Single tone test set up

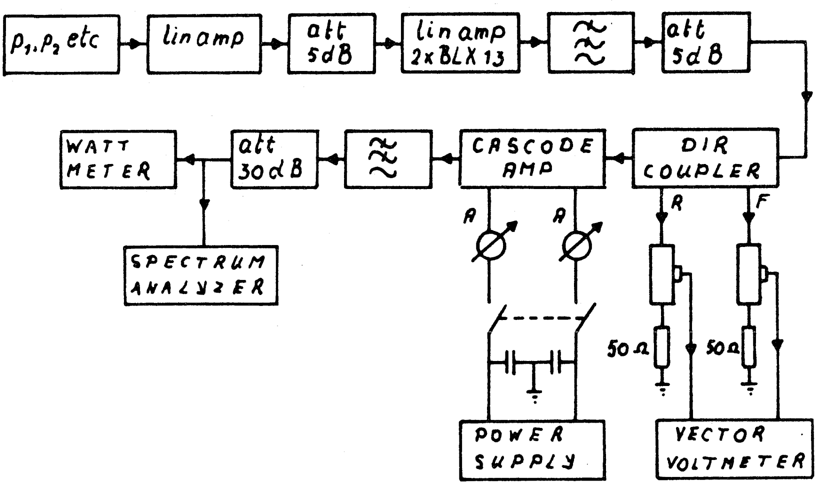

figure 6

6. References

1. M.J. Köppen - Wide-band (1.6-28MHz) driver stage for the SSB tube YL1230 equipped with two BLX13's in class A.
C.A.B. report ECO 7201.

2. A.H. Hilbers - On the design of H.F. wide-band power transformers.
C.A.B. report ECO 6907.

3. A.H. Hilbers - Design of H.F. wide-band power transformers, Part II.
C.A.B. report ECO 7213.

4. J. Mulder - A single stage wide-band linear power amplifier for 80 - 100W P.E.P. in the 1.6-28MHz frequency band equipped with two pieces BLX14.
C.A.B. report ECO 7114.

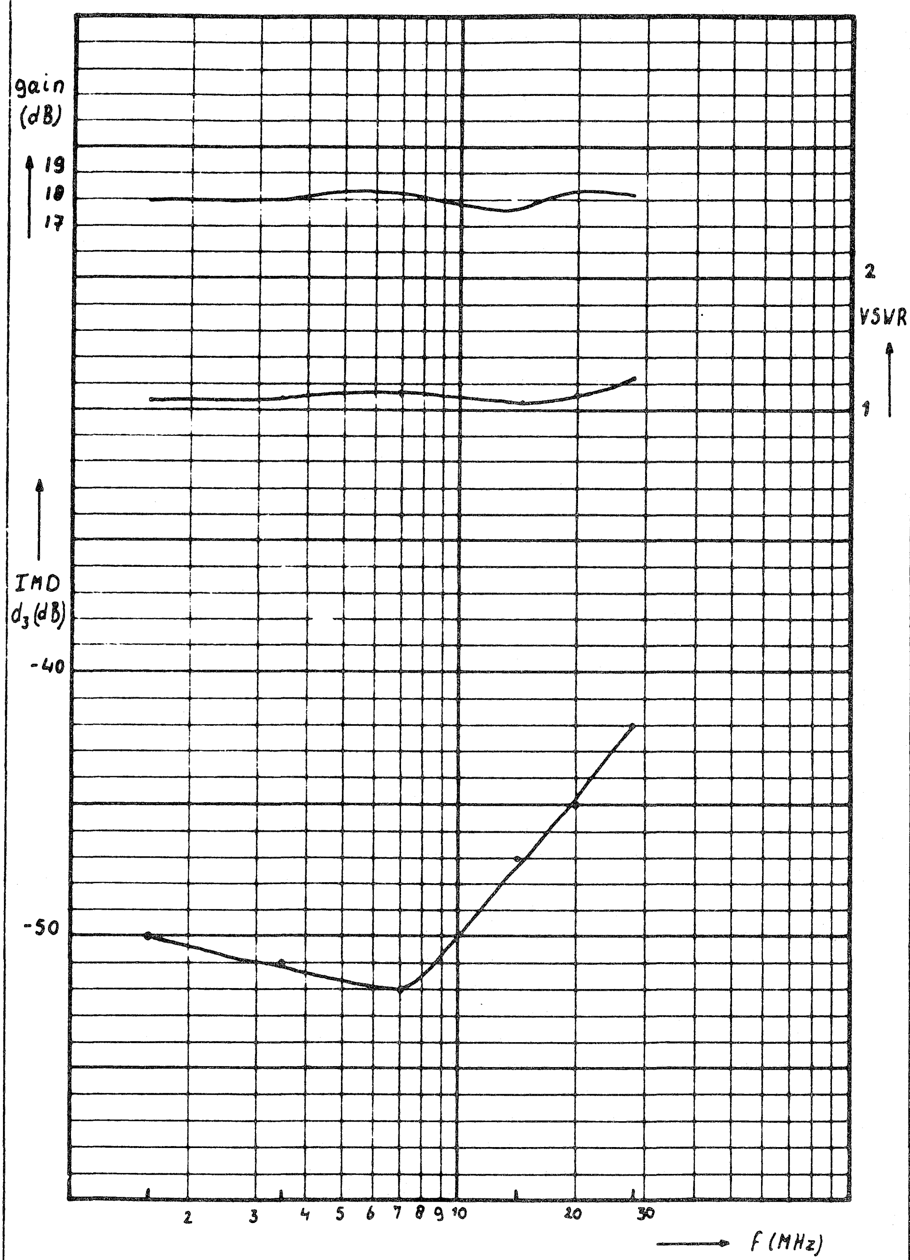

figure 7

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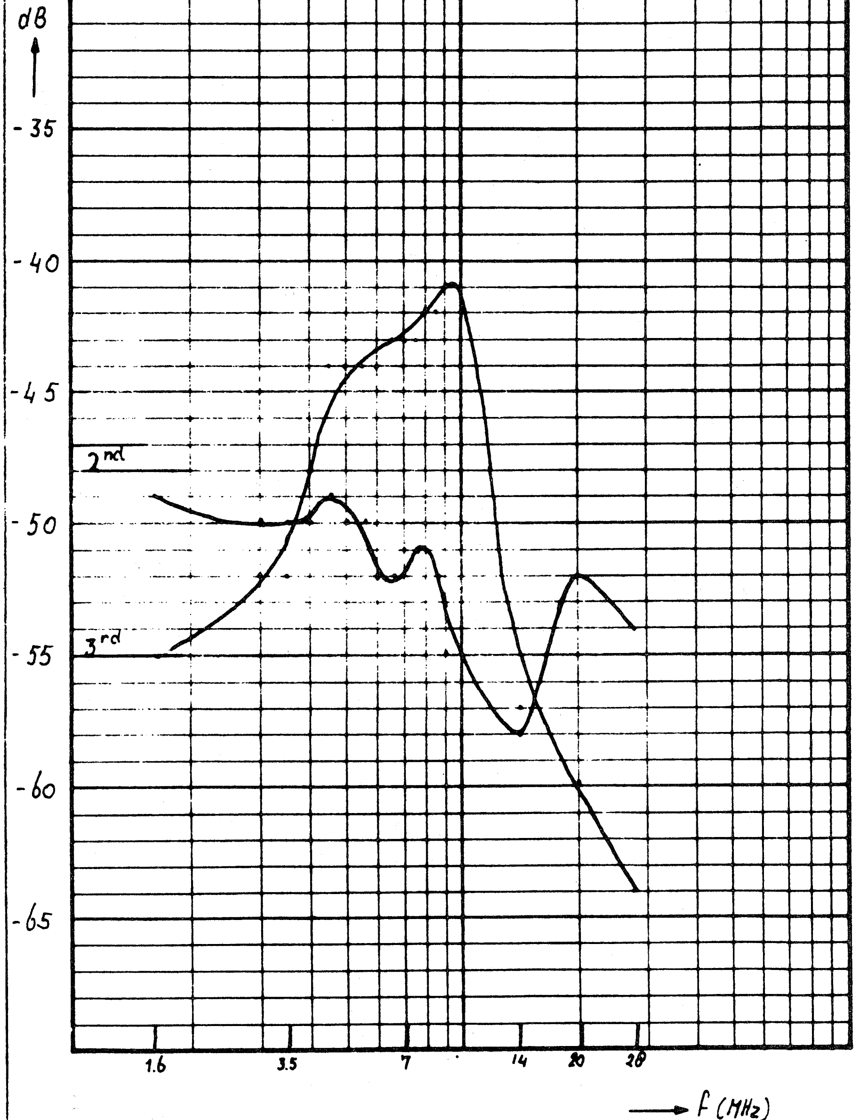

figure 8

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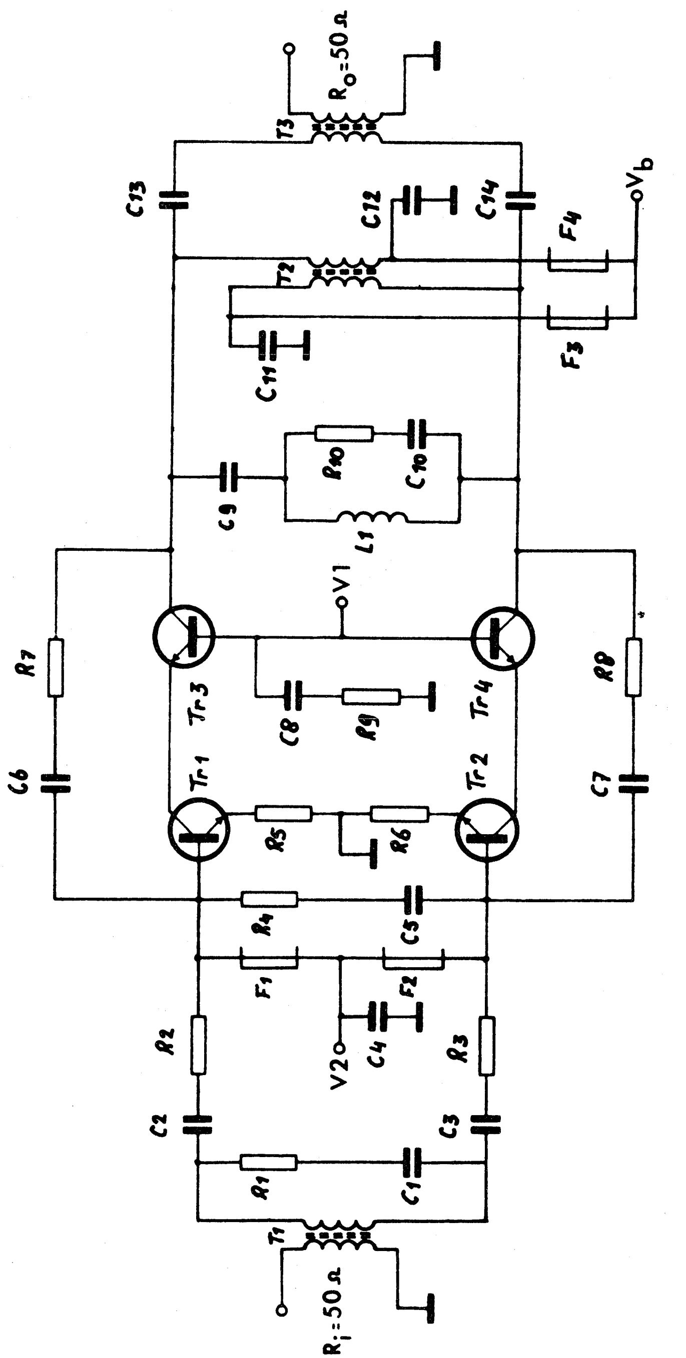

figure 9

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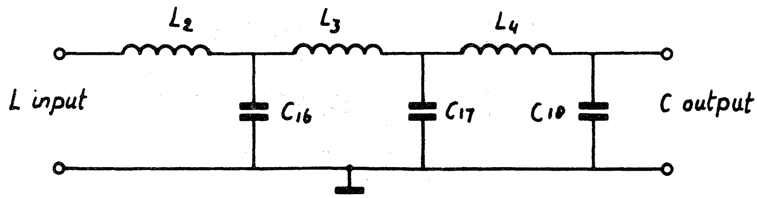
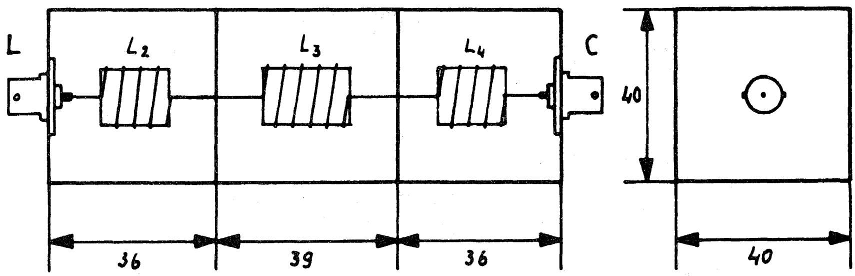
Low pass Chebyshev filter with L-input and C-output (Z_o = 50 Ohm)



Figuur 10a



(dimensions in mm)

figuur 10b

The filter is built in a box from single clad P.C. board and divided in three sections (L_2, L_3 and L_4) by pieces of double clad board, soldered between the walls.

The maximum input V.S.W.R. is 1.22 and the power loss at 28MHz is 3.7%.

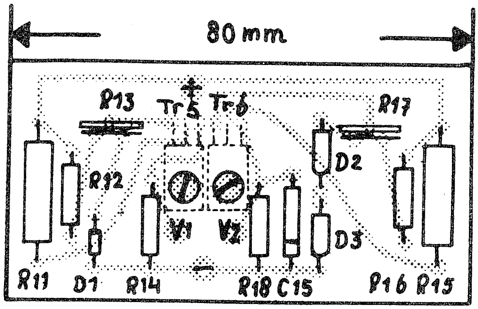
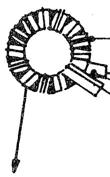


figure 11

T1

*) PTFE foil on the outside

figure 12



n_1: 20 turns of 0.6 mm enam Cu wire

n_2: 14 turns

PTFE foil *) with adhesive copper foil

4 mm wide — 2 mm wide

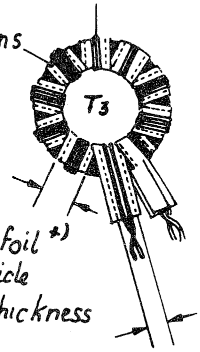
0.1 mm thickness — thickness appr 75 μ

Core: Fxc toroid, grade 4C6
14 x 9 x 5 mm³
code number:
4322 020 91020

figure 14

n_2: 16 turns of 2 x 0.6 mm enam Cu wire

n_1: 10 turns

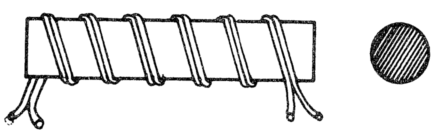


PTFE foil *)
6 mm wide
0.1 mm thickness

adhesive copper foil
4 mm wide
thickness appr 75 μ

Core: Fxc toroid, grade 4C6
23 x 14 x 7 mm³
code number:
4322 020 91070

T2: Part of antenna rod: 4311 020 55450



n_1, n_2: 6 turns of 2 x 1.1 mm enam Cu wire wound on a Fxc rod, grade 4A10
Dimensions (D x l): 10 x 50 mm²

fig 13

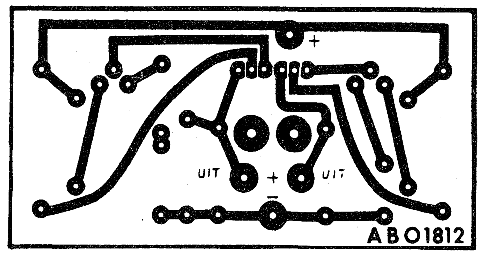


figure 15

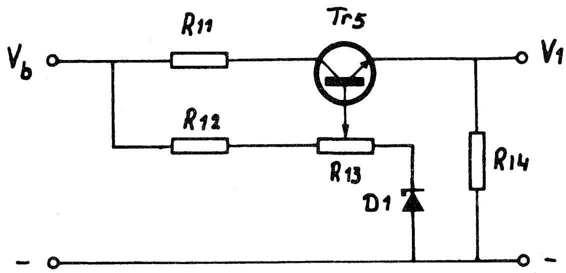


figure 16a

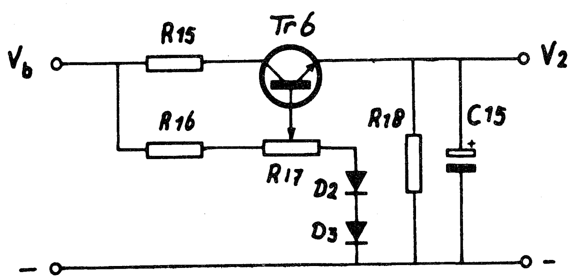


figure 16b

Parts list

Resistors

R_1	= 15 ohm	CR37 style	2322 212 13159
R_{2-3}	= 2x12 ohm in parallel	CR37 style	2322 212 13129
R_4	= 12 ohm	CR37 style	2322 212 13129
R_{5-6}	= 6x3.3 ohm in parallel	CR68 style	2322 214 13338
R_{7-8}	= 2x56 ohm in series	CR68 style	2322 214 13569
R_9	= 10 ohm	CR37 style	2322 212 13109
R_{10}	= 15 ohm	PR52 style	2322 192 31509
R_{11}	= 100 ohm	WR0617E	2322 330 22101
R_{12}	= 2.2Kohm	CR37 style	2322 212 13222
R_{13}	= 220 ohm potmeter miniature		2322 410 05002
R_{14}	= 1.8Kohm	CR37 style	2322 212 13182
R_{15}	= 120 ohm	WR0617E	2322 330 22121
R_{16}	= 3.3Kohm	CR37 style	2322 212 13332
R_{17}	= 220 ohm potmeter miniature		2322 410 05002
R_{18}	= 220 ohm	CR37 style	2322 212 13221

Capacitors

C_1	= 150pF "micropoco"	polystyr.	2222 427 41501
C_{2-3}	= 47nF "flat film"	polyester	2222 342 45473
C_4	= 100nF "flat film"	polyester	2222 342 45104
C_5	= 240pF "micropoco"	polystyr.	2222 427 42401
C_{6-7}	= 100nF "flat film"	polyester	2222 342 45104
C_8	= 2x470nF in par. "micropoco"	polystyr.	2222 426 44601
C_{9-10}	= 150pF "micropoco"	polystyr.	2222 427 41501
C_{11-12}	= 100nF "flat film"	polyester	2222 342 45104
C_{13-14}	= 10nF "flat film"	polyester	2222 342 45103
C_{15}	= 4.7 µF electrolytic	aluminium	2222 015 15478

C_{16} = 150pF "micropoco" polystyr. 2222 427 41501
in parallel with 12pF ceramic 2222 650 10129

C_{17} = 180pF "micropoco" polystyr. 2222 427 41801
in parallel with 8.2pF ceramic 2222 650 09828

C_{18} = 100pF "micropoco" polystyr. 2222 427 41007

Inductors

L_1 = 40nH, 3 turns of 0.6mm enam Cu wire
internal diameter 4.3mm, length 1.8mm

L_2 = 247nH, 5 turns of 1.0mm enam Cu wire,
internal diameter 11mm, length 10mm,
wound on hard-paper tube.

L_3 = 472nH, 7 turns of 1.0mm enam Cu wire,
internal diameter 12mm, length 12mm,
wound on hard-paper tube.

L_4 = 406nH, 7 turns of 1.0mm enam Cu wire,
internal diameter 11mm, length 13mm,
wound on hard-paper tube.

$F_{1-2-3-4}$ = Ferroxcube wide-band H.F. choke
grade 3B 4312 020 36640

Semiconductors

Tr_{1-2}	BLW60
Tr_{3-4}	ON495 (585BLY)
Tr_{5-6}	BD135
D_1	BZX79C15
D_{2-3}	BY206

Transformers

T_1	see Figure 12
T_2	see Figure 13
T_3	see Figure 14

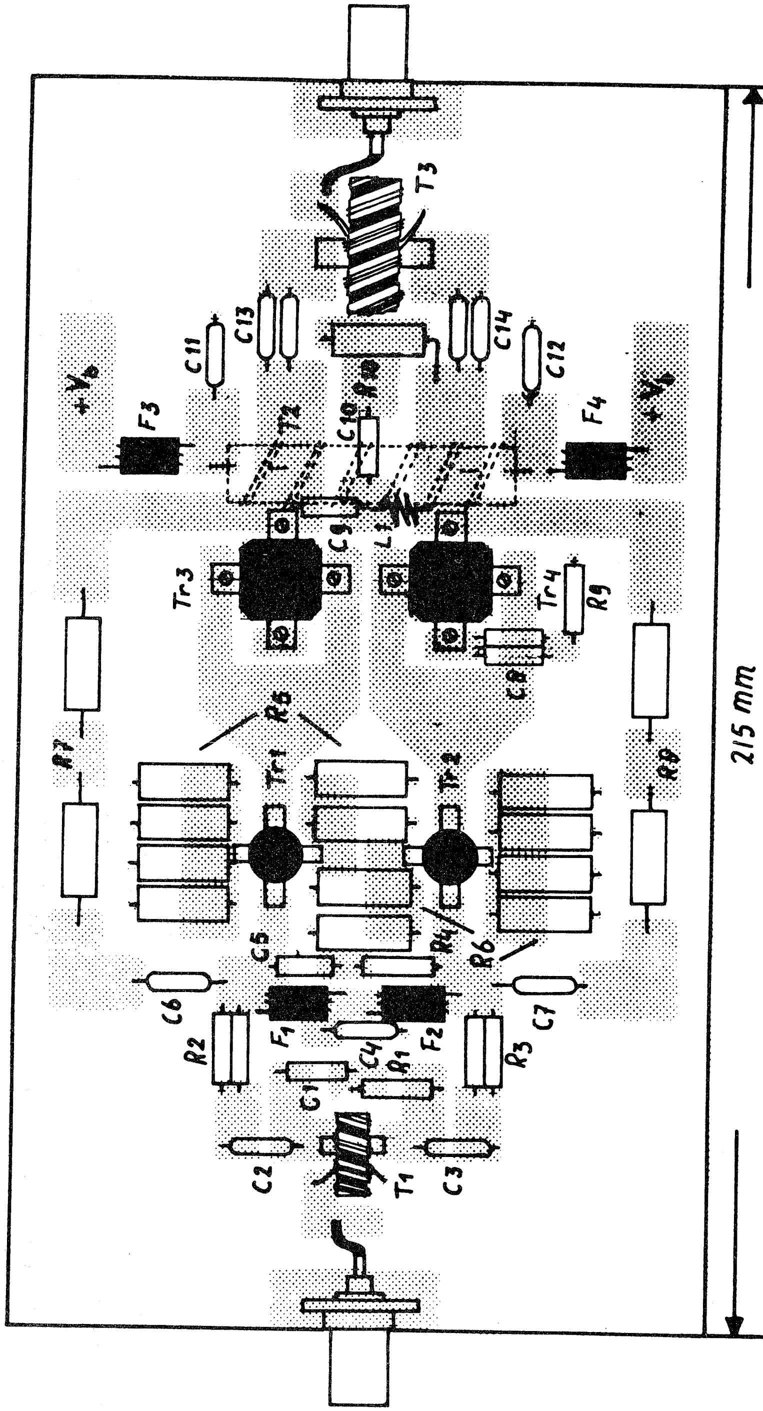

T3
C13
C11
C14
C12
F3
F4
Tr 3
Tr 4
R9
L1
C9
R7
R8
Tr1
Tr2
C5
R4
R6
C6
C7
R2
R3
F1
F2
C4
R1
C1
C2
C3
T1
215 mm

figure 17

PHILIPS

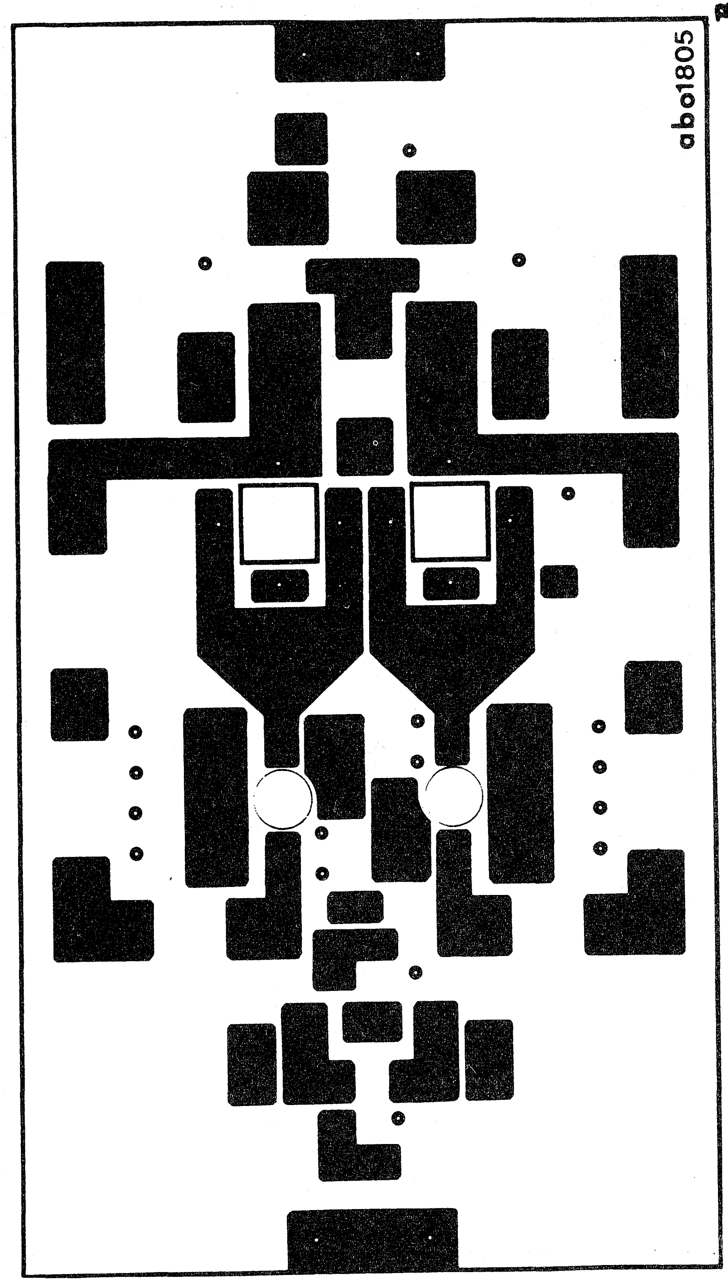
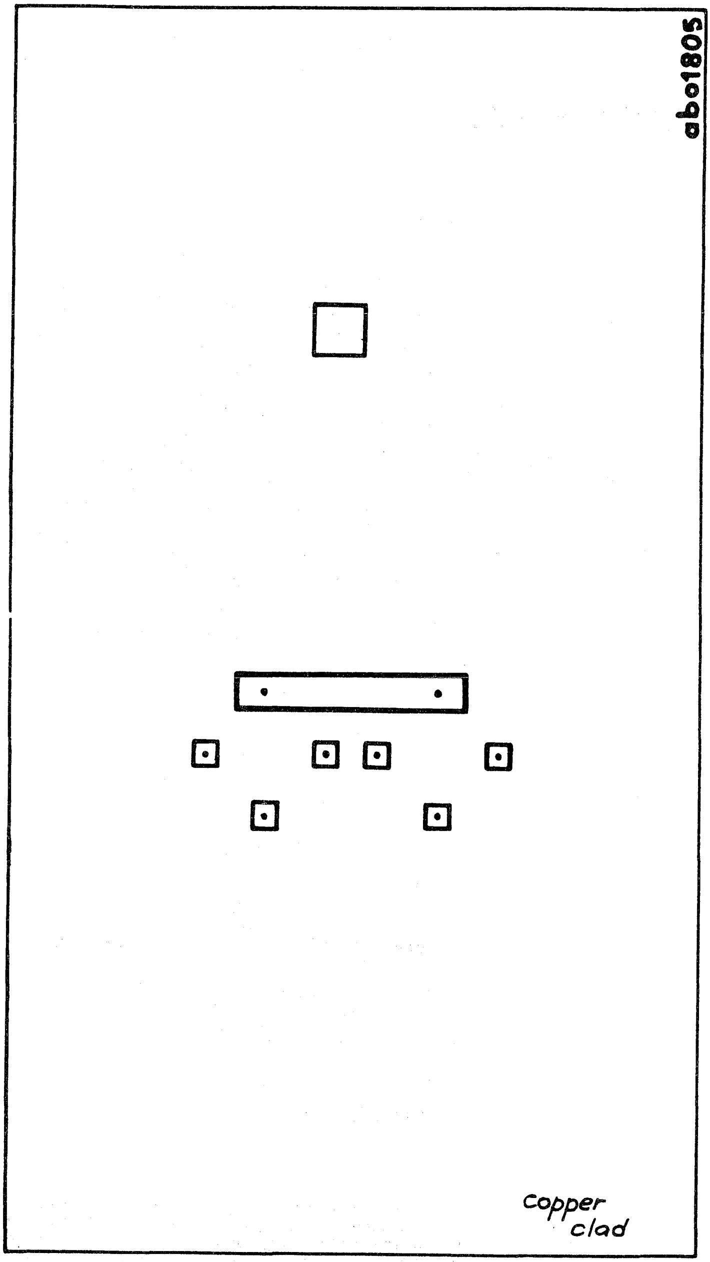


figure 18

PHILIPS

8222 031 76361

A



B

figure 19

negative

Electronic components and materials

Central application laboratory CAB Eindhoven - The Netherlands

Electronic
components
and materials

PHILIPS

laboratory report

number :MC08002   date : 1980 jun 16

title :A Single Stage Wideband (1.6MHz to 30MHz) Linear Amplifier For 400 Watts P.E.P. Using BLW96 Transistors.

author :J. Ling

Mullard

laboratory report

from the applications laboratory:

Mullard application laboratory,
Mitcham, England.

12 AUG. 1980

number:	MC08002	date:	1980 JUN 16
project:		pages:	; ;

title: A Single Stage Wideband (1.6MHz to 30MHz) Linear Amplifier For 400 Watts P.E.P. Using BLW96 Transistors.

author: J. LING

approved: A.J. REES

ABSTRACT

This report shows that a wideband power amplifier module providing up to 400 watts p.e.p. from 1.6MHz to 30MHz can be made using two BLW96 transistors.

Operating from a 50 volt supply rail the intermodulation distortion at 400 watts p.e.p. over the band is better than -26dB when loaded with a 50Ω wideband load.

The gain of the module is between 13.5dB and 16dB over the band and the input v.s.w.r. is better than 1.5:1 over the band.

For applications below 400 watts, e.g. 200W or 300W, improved intermodulation performance is obtainable.

ADVIES 1980-08-28 OCTROOI d.d.	X AV	GV		B		BL
OPGAVE 1980-08-12 MAMO d.d.	X AV	X GV	X EI	B		BL

JL/AJR/KIM

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MC08003

A SINGLE STAGE WIDEBAND (1.6MHZ TO 30MHZ) LINEAR AMPLIFIER FOR 400 WATTS P.E.P. USING BLW96 TRANSISTORS

1. INTRODUCTION

The design of this amplifier was to exploit the performance of the BLW96 in a wideband circuit to provide up to 400 watts p.e.p. from 1.6MHz to 30MHz.

2. CIRCUIT DESCRIPTION

Fig. 1 shows the circuit diagram.

Fig. 2 shows the circuit board layout.

Fig. 3 & 4 show the component layout.

It will be noted that baluns are used in input and output circuits. This improves the balance of collector currents over the frequency range, offsetting any capacitive unbalance in the impedance matching transformers.

Cross neutralisation has been found to be detrimental with these devices, the practical application of neutralisation results in excessive phase changes and results in higher level of intermodulation distortion over parts of the band.

The input matching network was computer designed as described in previous reports (1,2). The network is primarily intended to provide a constant overall gain with frequency and reasonably constant load impedance to a driver circuit.

In practice a capacitive centre tap (C_4, C_5) at the secondary of the input impedance matching transformer T_1 is desirable, which is earthed via a 2.2Ω resistor.

It is preferable in realising C_4, C_5, to employ three capacitors of 470pF, 470pF and 390pF in parallel for both, leaving the centre connection such that alternative connections of R_1 can be selected, as shown in Fig. 1. This technique allows the best compromise to be found for the intermodulation distortion peaks which typically occur between 14 and 20MHz, and which can be undesirably high at spot frequencies.
A separate, temperature compensated, bias control circuit is necessary to give adjustable constant class AB bias condition in the BLW96 transistors. The bias circuit diagram is shown in Fig. 1.

3. DESIGN DETAILS

To obtain the best performance from the amplifier over the complete range 1.6MHz to 30MHz, the quality of the matching transformers used is very important as accurate impedance transformation and low losses are essential. Low losses can be achieved by using toroids of 4C6 material. To obtain accurate matching it is essential to make transformers with very low leakage reactance at 30MHz combined with sufficient primary inductance at 1.6MHz (3,4).

3.1 The Output Transformer

The load impedance required by the transistors is 9.6Ω, collector to collector. A transformer is required to match to the 50Ω load. To handle the expected power of 400 watts the transformer must be wound using 2 4C6 cores each of 36 x 23 x 15mm. The best results were obtained using two separate transformers, each on one core, and connecting primary and secondary windings in parallel to give an acceptable value of leakage reactance at 30MHz.

The complete transformer, wound as detailed in the parts list had a secondary reactance (50Ω winding) of 200Ω at 1.6MHz (20μH) and a leakage reactance of 50Ω at F = 30MHz (265nH).

3.2 The Input Transformer

The design impedance of the gain correction network is 5.55Ω, so a transformer is required to match the input network to a 50Ω drive source.

The required drive input power assuming a minimum gain of 13dB is about 20 watts which requires two 4C6 toroids 14 x 9 x 5mm. Again, the best results were obtained by the parallel connection of two transformers each using a separate core.

The complete transformer, wound as detailed in the parts list had a primary reactance (50Ω winding) of 220Ω at 1.6MHz (22μH) and a leakage reactance of 50Ω at 30MHz (265nH).

3.3 The Bias Unit

The circuit uses a BD433 as the temperature sensor and a BD203 emitter follower. The unit can supply a maximum bias current of 800mA, dependent on the value of resistor in the collector of the BD203. The bias unit should be thermally connected to the amplifier by mounting on the heatsink close to the BLW96's. The collector load resistor (three 17 watt wire wound resistors in parallel) can be mounted separately on stand off insulators at any convenient part of the heatsink.

3.4 The Centre Tapped Choke

The collector d.c. supply to each BLW96 is fed via a centre tapped coil arrangement which consists of a ferrite aerial rod (4A10 grade or equivalent) and has a inductance of 4.6μH which forms part of the total collector load.

4. PERFORMANCE OF THE AMPLIFIER

4.1 General

The measured performance of the amplifier i.e. intermodulation distortion, gain and input v.s.w.r. is given in graphical form in Figs. 5 to 8. A water cooled heatsink was used for all measurements, and a wideband 50Ω load was used in conjunction with a thermal power measurement system. The p.e.p. was assumed to be 2 x the r.m.s. power indicated, ignoring the harmonic content of the output signal.

4.2 Harmonic Output

The amplifier was driven with c.w. signals at specific frequencies to 400W, 300W and 200W. The driving signal had a harmonic content lower than -45dB. Harmonic component measurements are shown below in Table 1.

TABLE 1

HARMONIC OUTPUT V. FREQUENCY

Load Power W	Test Frequency MHz	Harmonic Content dB f_2	f_3	f_4	f_5	f_6	f_7	f_8	f_9	f_{10}
400	1.6	-31	-20	-37	-34	-39	-44	-43	-46	-52
	3.5	-36	-21	-45	-34	-63	-40	-48	-49	-47
	7	-38	-17	-47	-29	-45	-34	-64	-64	
	10	-45	-15	-45	-29	-50				
	14	-39	-16	-45	-46					
	20	-48	-24	-50	-44					
	28	-44	-41							
300	1.6	-31	-21	-39	-33	-40	-47	-43	-46	-52
	3.5	-36	-23	-45	-32	-65	-42	-48	-52	-49
	7	-40	-22	-52	-30	-48	-38	-65	-64	
	10	-48	-16	-45	-29	-50				
	14	-42	-18	-48	-47					
	20	-42	-25	-50	-46					
	28	-45	-42							
200	1.6	-32	-22	-40	-33	-42	-49	-44	-47	-55
	3.5	-35	-24	-46	-33	-65	-43	-50	-53	-50
	7	-40	-23	-62	-30	-59	-39	-65	-64	
	10	-50	-19	-45	-30	-50				
	14	-44	-19	-47	-47					
	20	-38	-27	-49	-47					
	28	-44	-42							

4.3 Collector Efficiency

The collector efficiency varies over the band, and as the output matching was designed for 400 watts, reduced efficiency is apparent at reduced levels of output. Typical efficiencies are:-

(a) Two Tone Signals Between 1.6MHz and 30MHz

P.E.P. load W	Best Collector Efficiency %	Worst Collector Efficiency %
400	50	37.7
300	44	32

Taking the worst case and assuming equal sharing and zero circuit losses each transistor would dissipate 170 watts.

(b) C.W. Signals Between 1.6MHz and 30MHz

P load W	Best Collector Efficiency %	Worst Collector Efficiency %
400	65.5	47.5

Taking the worst case and assuming equal sharing and zero circuit losses each transistor would dissipate 220 watts.

The published r.f. thermal resistance of the BLW96 is 0.65 K/W (junction-heatsink) and the maximum permissible junction temperature is 200°C. Therefore the maximum possible heatsink temperature for continuous C.W. operation is 200 - (220 x 0.65) = 57°C.

5. BIAS ADJUSTMENT

The zero signal collector currents should be adjusted by the bias potentiometer to 2 x 100mA when operating from a 50 volt supply.

6. CONCLUSIONS

6.1 A single stage amplifier using 2 x BLW96's in class AB can deliver up to 400 watss p.e.p. with intermodulation distortion better than -26dB over the band 1.6MHz to 30MHz.

6.2 Measurements indicate that when used under C.W. conditions at the frequencies where the maximum collector dissipations occurs, the devices should be capable of operation with heatsink temperatures up to 57°C.

7. REFERENCES

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   C.A.B. Report E.C.O.7505

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4. A.H. Hilbers — Design of H.F. Wideband Transformers Part II.
   C.A.B. Report E.C.O.7213

APPENDIX I

PARTS LIST Figs. 1a, 1b, 1c, 1d

1a Transistors and Resistors

TR_1 TR_2	BLW96	
TR_3	BD433	
TR_4	BD203	

Component	Value	Type		Philips Code No.		
R_1	= 2.2Ω	CR 37 ±5%		2322	212	13228
R_2,R_3	= 18Ω	PR 37 ±5%		2322	191	31809
R_4,R_5	= 2 x 12Ω	PR 37 ±5%	in parallel	2322	191	31209
	2 x 15Ω	PR 37 ±5%		2322	191	31509
R_6	= 1k5Ω	PR 37 ±5%		2322	191	31502
R_7	= 3 x 180Ω	EH 15 ±5%	in parallel	2306	330	03181
R_8	= 3.3Ω ajustable TPW22			2322	011	02338
R_9	= 22Ω	CR 37 ±5%		2322	212	13229

1b Capacitors

Component	Value		Philips Code No.		
C_1	= 10nF polyester ±20%		2222	342	44103
C_2	= 60pF trimmer		2222	809	08003
C_3	= 330pF polystyrene ±1%	in parallel	2222	426	43301
	300pF " ±1%		2222	426	43001
C_4,C_5	= 2 x 470pF polystyrene ±1%	in parallel	2222	426	44701
	1 x 390pF polystyrene ±1%		2222	426	43901
C_6,C_7	= 2 x 1000pF polystyrene ±1%	in parallel	2222	426	44102
	1 x 820pF polystyrene ±1%		2222	426	48201
C_8	= 3 x 100nF polyester ±20%	in parallel	2222	342	44104

C_9, C_{10}	=	2 x 47pF ceramic ±2%	in parallel	2222	650	34479
		2 x 56pF ceramic ±2%		2222	650	34569
C_{11}, C_{12}	=	5 x 10nF polyester ±20%	in parallel	2222	342	44103
C_{13}, C_{14}	=	3 x 100nF polyester ±20%	in parallel	2222	342	44104
C_{15}	=	33pF ceramic ±2%		2222	650	34339
C_{16}, C_{17}	=	3.3µF ±10%		2222	344	21335
C_{18}, C_{19}	=	220µF 4V electrolytic		2222	016	2221
C_{20}, C_{21}	=	100nF polyester ±20%		2222	342	44104
C_{22}	=	220µF 10V electrolytic		2222	016	4221

1c **Inductors**

L_1, L_2 = 12.9nH consists of $\frac{3}{4}$ turn of 1.3mm cu wire 5mm diameter

L_3, L_4 = 21nH consists of 1 turn of 1.3mm cu wire 7.5mm diameter

Ch_1, Ch_2 = 2.5 turns through 6 hole ferrite bead grade 3B Code No. 4312 020 31500

Ch_3, Ch_4 = 3 parallel turns through 6 hole bead grade 3B Code No. 4312 020 31500

1d **Transformers**

T_1 = Two transformers with parallel connected primary and secondary windings.

Each transformer wound on 4C6 toroids 14 x 9 x 15mm. Code No. 4322 020 91020.

Primary winding consists of 10 turns of copper tape approximately 1.5mm. wide.

Secondary winding consists of 30 turns of 0.45mm. chain enamelled cu. wire.

Windings separated by a layer of p.t.f.e. tape approximately .025mm. thick.

T_2 = 4 turns of 2 x 1.0mm. enamelled cu. wire twisted, on a 50mm. length of 4A10 or equivalent grade aerial rod 10mm. diameter. Code No. 4311 020 55390.

T_3 = Two transformers with parallel connected primary and secondary windings.

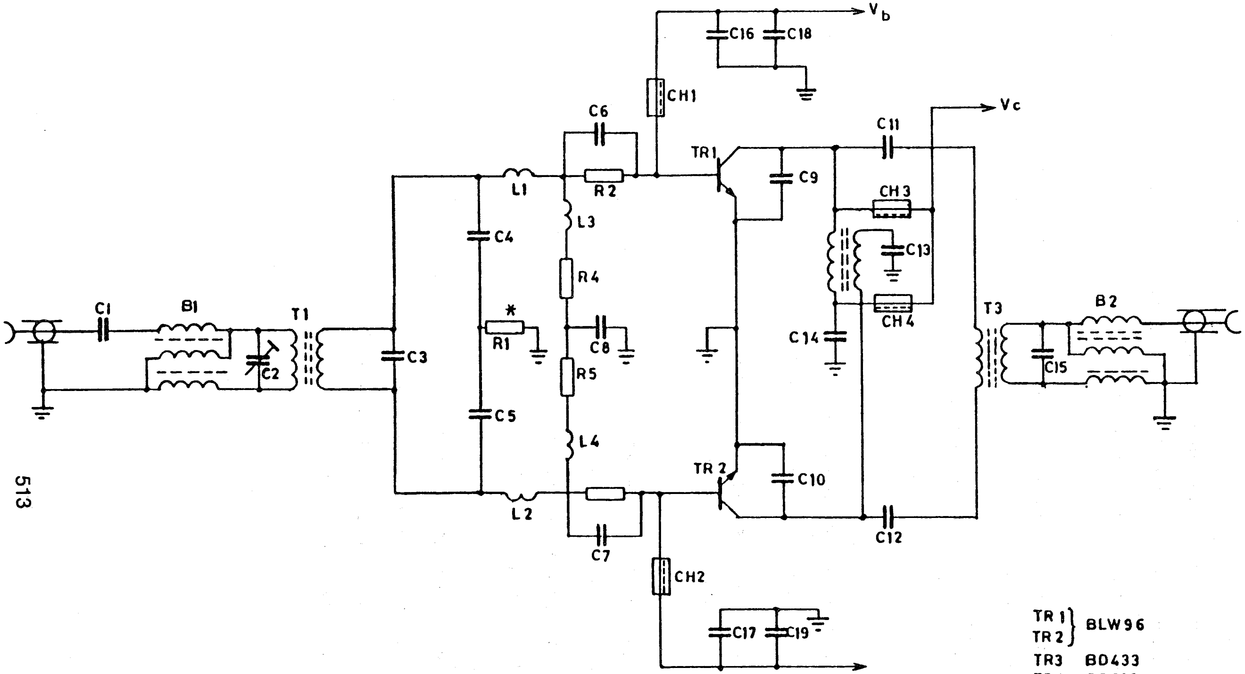
Each transformer wound on 4C6 toroids 36 x 23 x 15mm. Code No. 4322 020 91090.

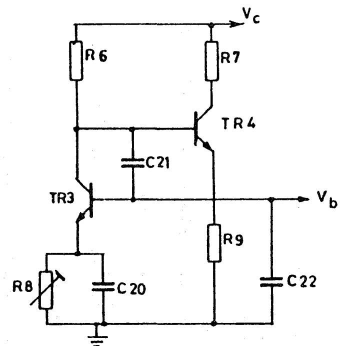
Primary winding consists of 6 turns copper tape 8.0mm. wide. Secondary winding consists of 14 turns of 4 x 0.5mm. diameter Cu. wire in parallel.

Windings separated by a layer of p.t.f.e. approximately .025mm. thick.

B_1 11 turns of 50Ω co-axial cable approximately 3mm. external diameter with p.t.f.e. dielectric and 11 turns of 0.5mm. enamelled cu. wire wound on 4C6 toroid 23 x 14 x 7mm. Code No. 4322 020 91070.

B_2 8 turns of 50Ω co-axial cable approximately 4mm. external diameter with p.t.f.e. dielectric and 8 turns of 1mm. enamelled cu. wire wound on 2 4C6 toroids 36 x 23 x 15mm. Code No. 4322 020 91090.





BIAS UNIT

TR1 } BLW96
TR2 }
TR3 BD433
TR4 BD203
* R1 SEE NOTE IN PARA.2.
L1,L2, ¾ turn 1·3mm Cu wire 5 mm dia.
L3 L4 1 turn 1·3mm. Cu wire 7·5 mm dia.

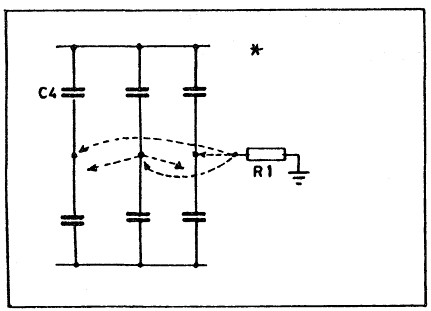


FIG. 1. BLW96 POWER AMPLIFIER 1·6 MHz–30MHz, 400 W. P.E.P

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DESCRIPTION				DRG. NO.				DATE	
ISSUE	1	2	3	4	5	6	7	8	9
DATE									

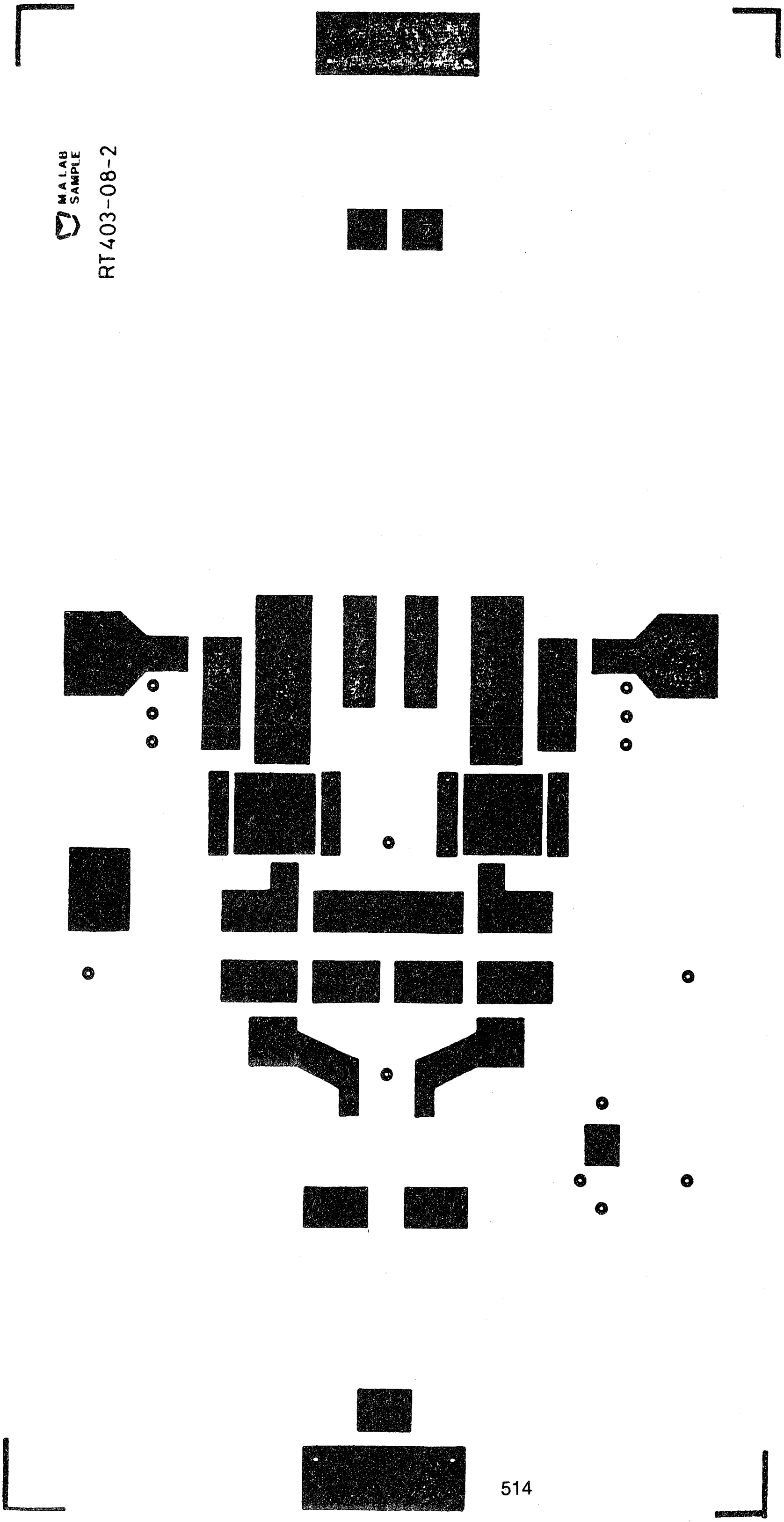


Fig. 2. BLW96 Power Amplifier Circuit Board

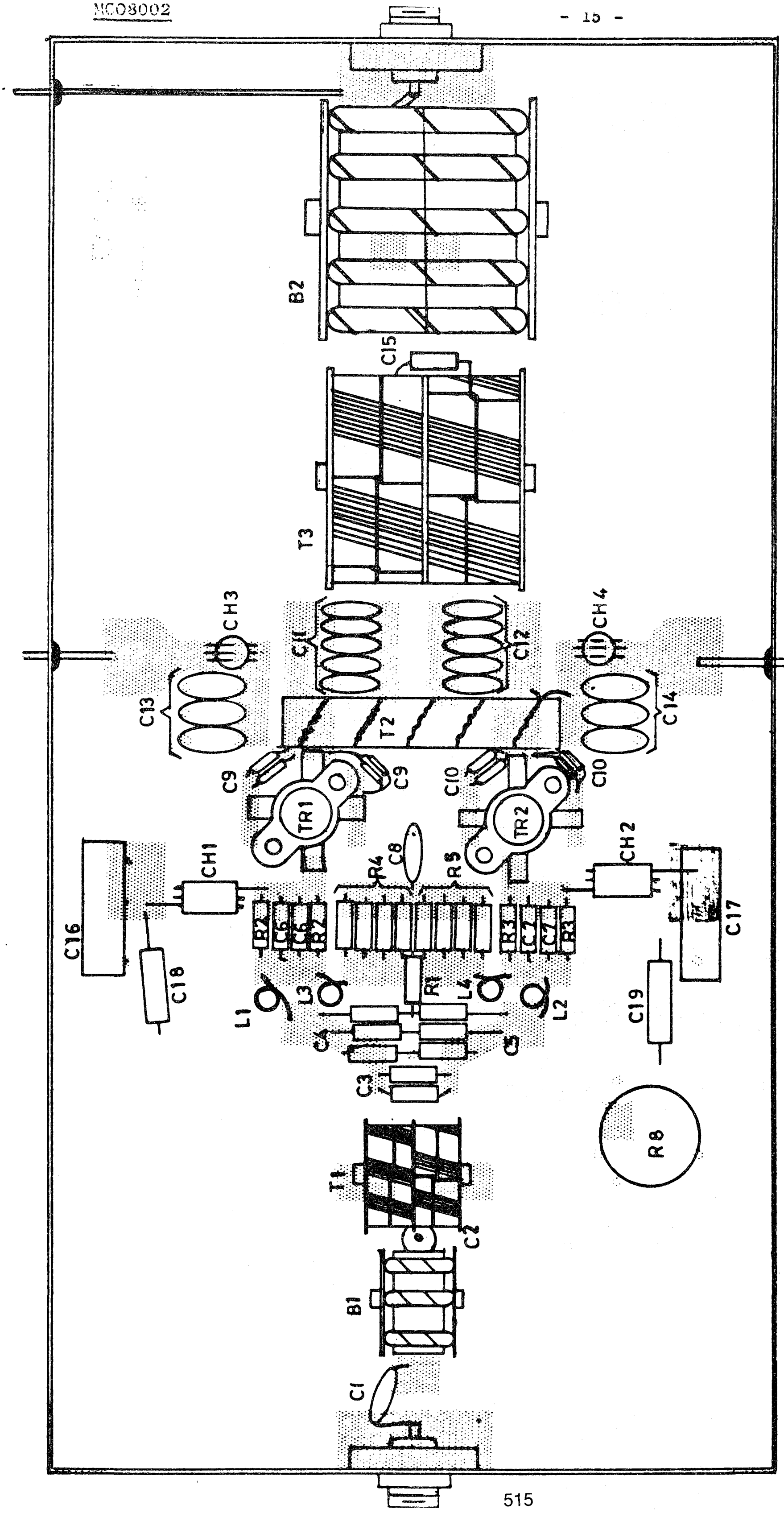
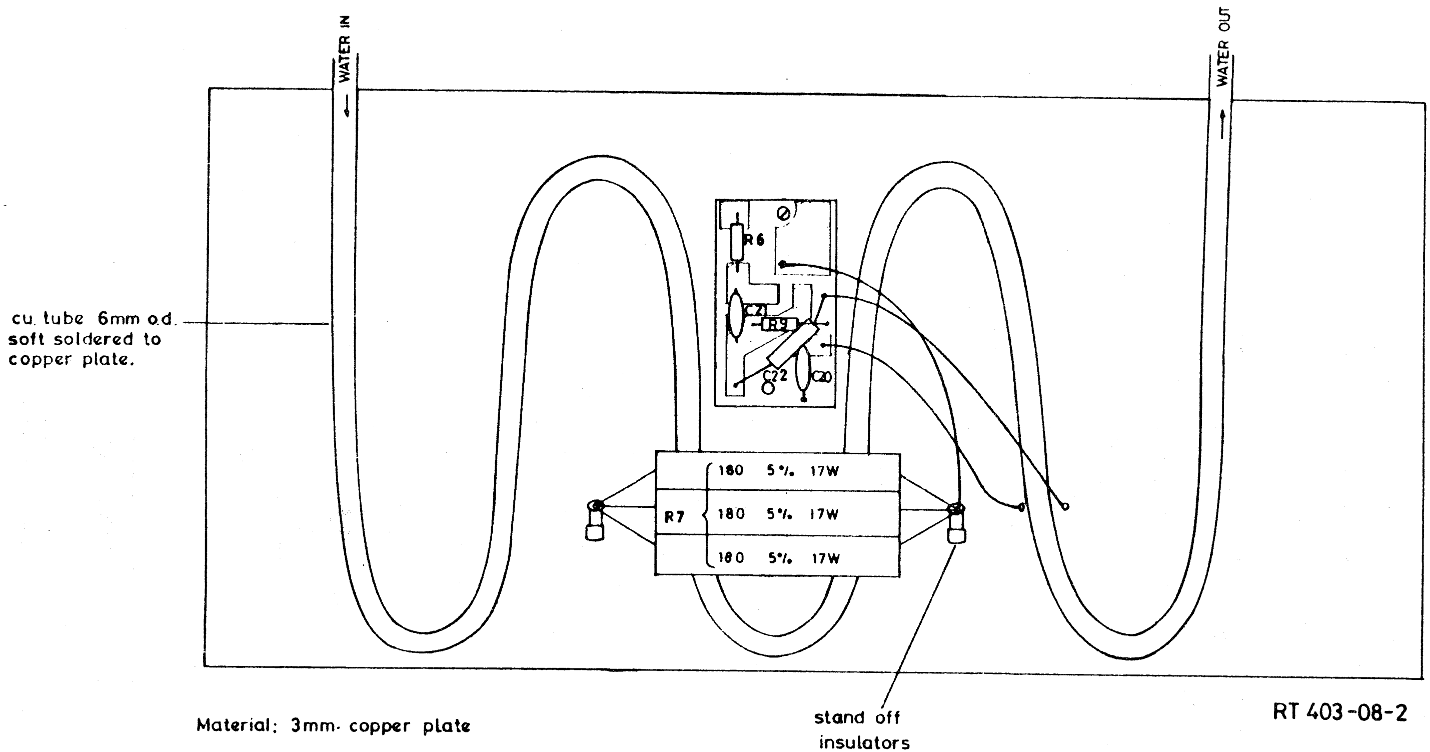
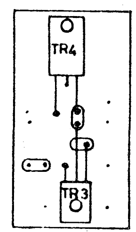


Fig.3. BLW96 Power Amplifier Component Layout





BIAS UNIT BOARD

Fig. 4. BLW96 Power Amplifier Component Layout. Underside of Heatsink.

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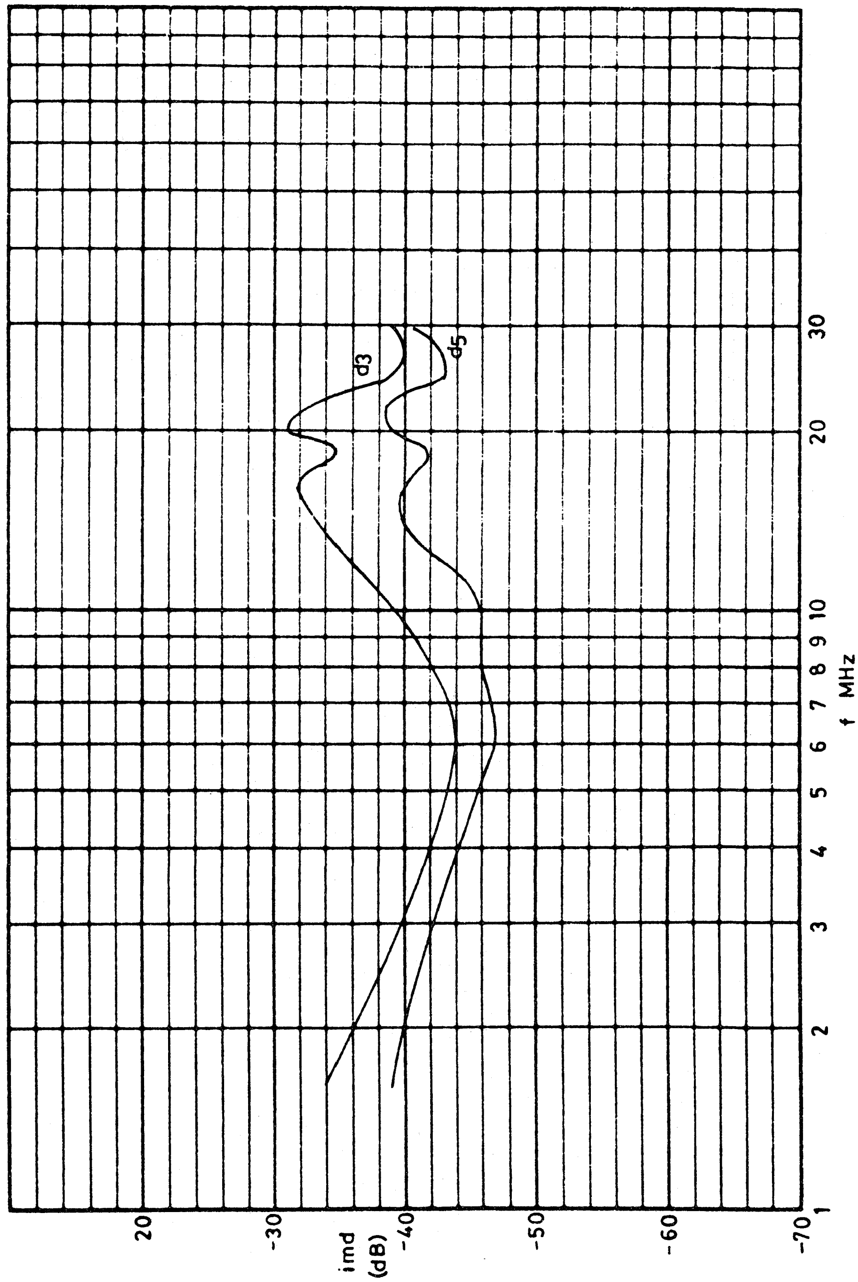
FIG. 5. INTERMODULATION PERFORMANCE OF BLW96 AMPLIFIER AT 200W PEP.



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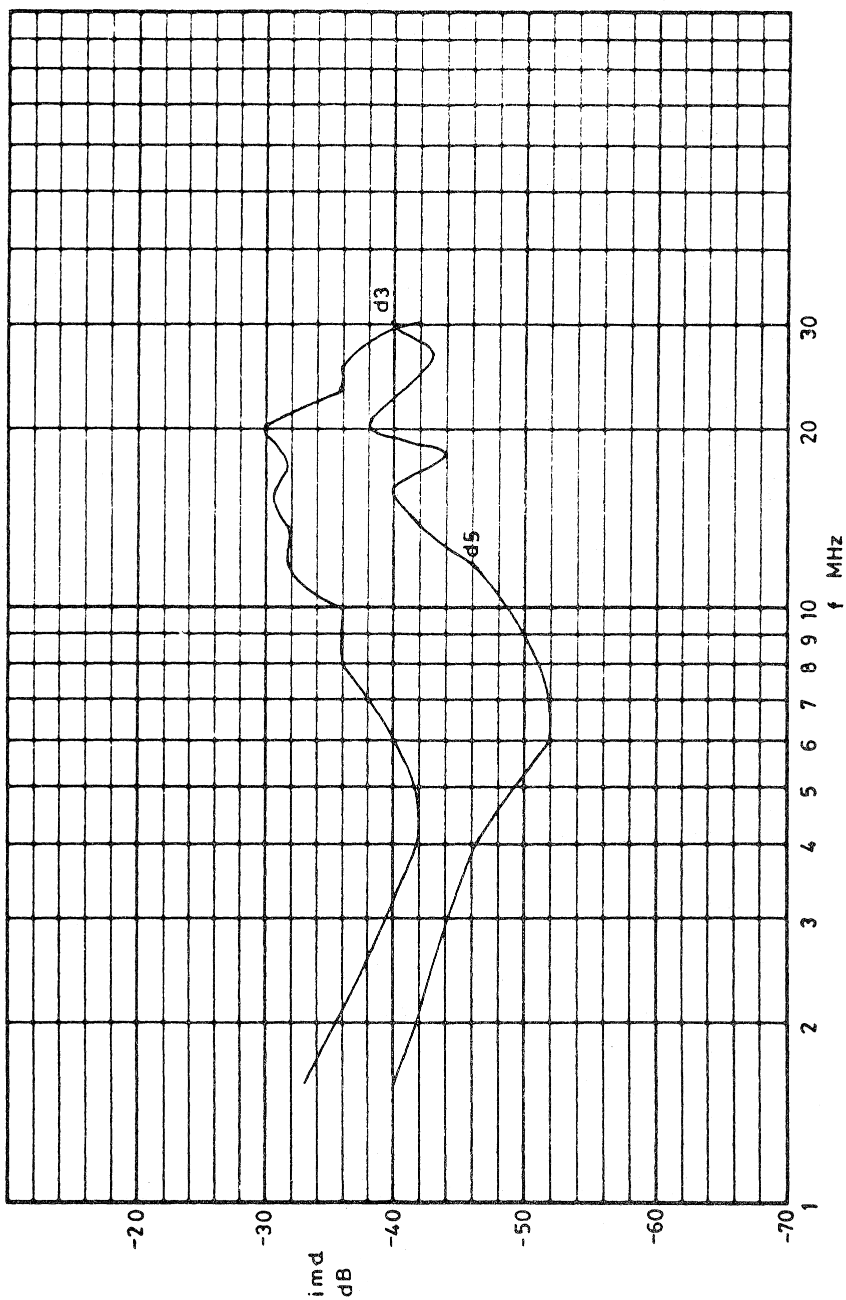

FIG. 6. INTERMODULATION PERFORMANCE OF BLW96 AMPLIFIER AT 300W PEP

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FIG. 7. INTERMODULATION PERFORMANCE OF BLW96 AMPLIFIER AT 400W P.E.P

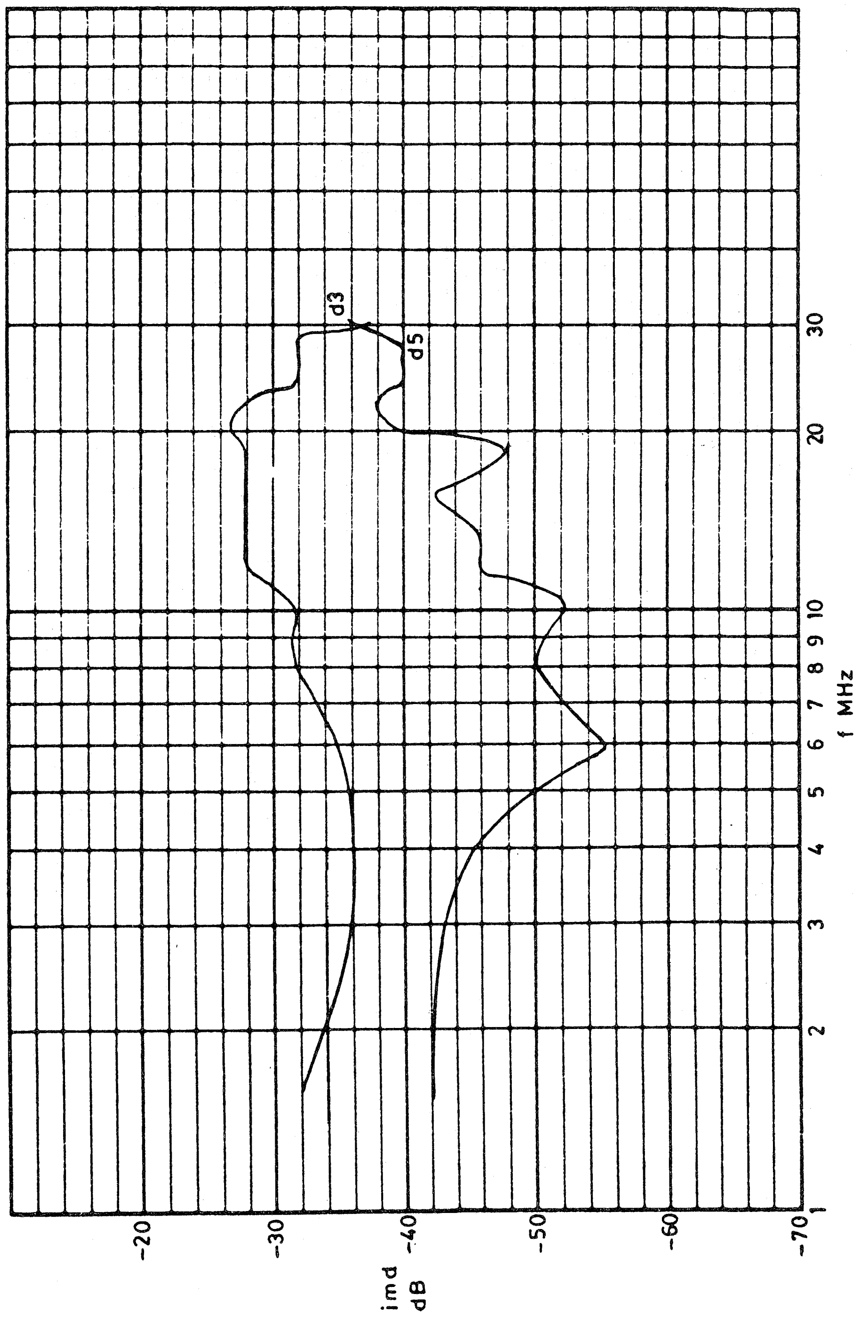

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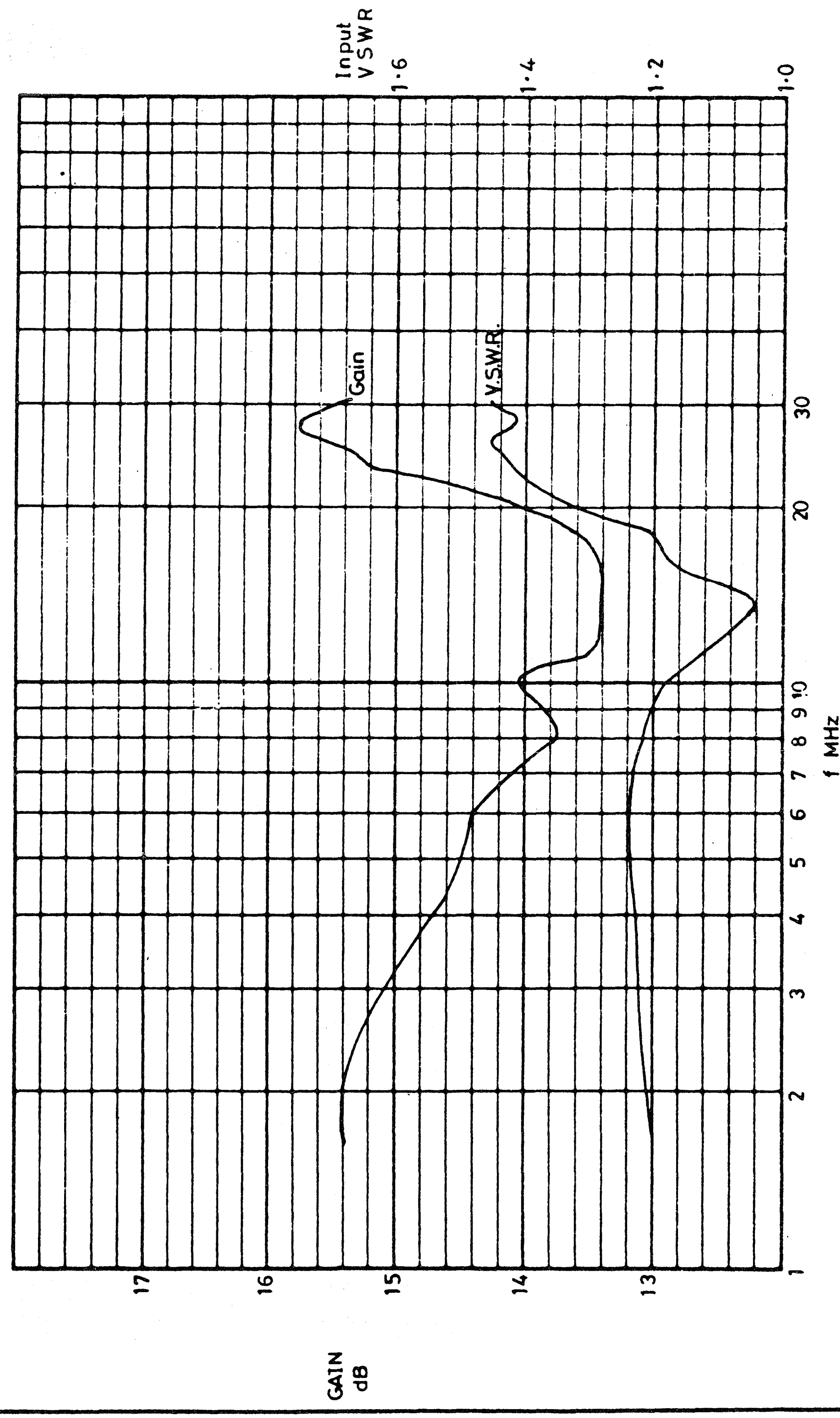
FIG.8. GAIN AND INPUT V.S.W.R. OF BLW96 AMPLIFIER AT 400W C.W.



Electronic
components
and materials

PHILIPS

laboratory report

number : MC08003 date : 1980 jul 01

title : A Single Stage Wideband (1.6 to 30MHz) Linear Amplifier For 25 Watts P-E-P Using BLW50F Transistors In Class A.

author : J. Ling

Mullard

laboratory report

from the applications laboratory:

Mullard application laboratory,
Mitcham, England.

12 AUG. 1980

number:	MC08003	date:	1980 JUL 01
project:		pages:	; ;
title:	A Single Stage Wideband (1.6 to 30MHz) Linear Amplifier For 25 Watts P-E-P Using BLW50F Transistors In Class A.		
author:	J. LING	approved:	A.J. REES

ABSTRACT

This driver stage will provide up to 25 watts p-e-p from 1.6 to 30MHz with intermodulation distortion level of -40dB into a matched 50Ω load. The overall gain is approximately 16dB ±0.5dB.

The amplifier can be used to drive a linear amplifier using two BLW96 to 400 watts p-e-p.

ADVIES OCTROOI d.d. 1980-08-28	xAV	GV		B		BL
OPGAVE MAMO d.d. 1980-08-12	xAV	xGV	xEI	B		BL

JL/AJR/KIM

MC08003

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MC08003

A SINGLE STAGE WIDEBAND (1.6 TO 30MHZ) LINEAR AMPLIFIER FOR 25 WATTS P-E-P USING BLW50F TRANSISTORS IN CLASS A

1. INTRODUCTION

This amplifier was developed to drive a 400 watt linear amplifier using BLW96 transistors in class AB and to provide data for future development of a two stage amplifier for the same power with both driver and p.a. stages working from a nominal 50 volt supply.

The driver amplifier described herein was designed to give 25 watts p-e-p into a 50Ω load sufficient to drive the amplifier previously reported[1]. Class A operation is necessary because the required intermodulation distortion of the driver circuit alone should be substantially less than that of the p.a., the driver circuit distortion being somewhat degraded when working into the non-linear input impedance of the p.a. A target figure of intermodulation products <-40dB over the frequency band 1.6 to 30MHz was therefore required.

As with the BLW96 p.a., 50Ω input and output impedances were assumed initially, but direct inter-stage impedance matching is used in the two stage amplifier which will be reported subsequently.

2. GENERAL CIRCUIT DESCRIPTION

The circuit diagram of the amplifier is given in Figure 1, and follows conventional class A push-pull linear h.f. amplifier design practice.

Some shunt and series feedback is applied: the shunt feedback resistors are also used to determine bias conditions.

The design and construction of the broadband input and output impedance matching transformers is critical in overall amplifier performance.

Details of these and other construction features are given below.

3. DESIGN AND CONSTRUCTION

3.1 Bias Condition

The design calculations for this amplifier in general repeat a method previously reported by Köppen[2]

In the design of linear class A amplifiers, in order to arrive at the best intermodulation performance, a suitable starting point is to assume maximum bias current and voltage permitted by the maximum anticipated heatsink temperature.

In this design, therefore, a practical maximum operating heatsink temperature of 70°C was assumed. At this heatsink temperature, the practical bias conditions for the BLW50F of V_{CE} = 44V, I_C = 1A are permissible.

It may be noted that the BLW96 amplifier[1] shows a worst case gain of about 13.5dB and a worst case v.s.w.r. of about 1.45 (these two conditions do not occur at the same frequency however). But, conservatively, we may say that a suitable driver should be capable of about 18.7 watts in a 50Ω load.

In addition, if this power (18.7 watts) is to be available at intermodulation products d3, d5, <-40dB, experience has shown that the maximum available power from the amplifier (class A) into a linear resistive load should approach 35 watts, at which power intermodulation products will of course exceed -40dB.

Further, with the circuit configuration chosen (Figure 1), it will be seen that appreciable power is dissipated in the shunt feedback resistors.

We expect about 15% of the available power will be so dissipated; therefore the maximum available power requirements becomes about 40 watts from the transistors.

The projected initial bias conditions, V_{CE} = 44V, I_C = 1A correspond to a collector dissipation of 44 watts per transistor. Under class operation the maximum output power would then be 22 watts per transistor, 44 watts total.

There is therefore some margin compared with the estimated capability requirement of 35 watts and this is reasonable, in part because some further degradation of intermodulation performance may be expected when directly driving into the somewhat non-linear input impedance of the p.a., instead of the linear, non reactive design load (50Ω).

Having chosen V_{CE} = 44V a further voltage drop of 2V may be assumed across the external emitter resistor for bias stabilisation.

3.2 Components and Layout

Figures 2 and 3 give details of the circuit board and component layout. The parts list in the Appendix includes details of wound components including the two wideband matching transformers.

A suitable heatsink must be provided, sufficient to limit the junction temperature to <200°C under d.c. (bias) conditions.

The total current required at 46V is about 2.25A, which includes base bias current flowing in the feedback resistors.

The total dissipation of the BLW50F transistors is, however approximately 44 x 2 = 88 watts.

The feedback resistors and external emitter resistors together dissipate a further 14 watts approximately.

If these resistors are mounted so that they do not contribute to the heatsink surface temperature close to the transistors, then the required heatsink thermal resistance, θ_{h-amb}, which determines a heatsink temperature of 70°C in a 25°C laboratory ambient, for example is given by:-

$$\theta_{h-amb} \leq \frac{70-25}{88}$$

$$\text{i.e. } <0.51^{\circ}\text{C/W}$$

To obtain the best compromise between input v.s.w.r. and gain, particularly between 26 to 30MHz, C_2 is made adjustable; this accommodates the inevitable spread of leakage reactance of the input transformer T_1.

Similarly, allowing for leakage reactance spreads in T_2, C_3 is also made adjustable.

4. AMPLIFIER PERFORMANCE

Figures 4 and 5 show the measured amplifier performance.

Figure 4 shows 3rd and 5th order intermodulation products relative to the amplitude of each tone (12.5 watts) under standard 2 tone drive conditions.

Figure 5 shows amplifier gain and input v.s.w.r. under single tone drive at P_{load} = 25 watts.

It is noted that d_3 is substantially <-40dB over the frequency range of interest; the minimum gain and worst case v.s.w.r. are 15.7dB, 1.36 respectively, both occurring at the lower frequencies.

5. CONCLUSIONS

A wideband linear h.f. amplifier has been designed and constructed, using BLW50F transistors in class A which can give up to 25 watts p-e-p over the band 1.6 to 30MHz with 3rd and 5th order intermodulation better than -40dB between 2 and 28MHz.

Amplifier gain and input v.s.w.r. are better than 15.7dB and 1.4:1 respectively over the frequency range. The amplifier should be suitable to drive a further linear amplifier using BLW96's in class AB to 400W p-e-p.

6. REFERENCES

1. J. Ling. A Single Stage (1.6 to 30MHz) Linear Amplifier For 400 Watts P-E-P Using BLW96 Transistors. Mullard Application Laboratory Report No. MC08002.

2. M.J. Köppen. A Single Stage Wideband (1.6 to 28MHz) S.S.B. Driver Module With BLY92A And BLX13 Operating In Class A.
Philips Central Application Laboratory. ECO7113.

APPENDIX

PARTS LIST

RESISTORS

R_1, R_2	= 2x200Ω ±5% in series	Philips type	PR52	2322	192	32001
R_3, R_4	= 5x12Ω ±5% in parallel	" "	CR25	2322	211	13129
R_5	= 6.8Ω ±5%	" "	CR25	2322	211	13688
R_6	= 22Ω ±5%	" "	CR37	2322	211	13229
R_7	= 15Ω ±5%	" "	CR25	2322	211	13159
R_8	= 3R3Ω adjustable		TPW22	2322		

CAPACITORS

C_1, C_4	10nF ±20% polyester	Philips type	352	2222	352	44103
C_2, C_3	60pF trimmers	" "	809	2222	809	08003
C_5	22nF ±20% polyester	" "	352	2222	352	44223
C_6, C_7	2x47nF ±20% polyester	" "	352	2222	352	44473

TRANSFORMERS

T_1 = 1:1.5 turns ratio.

Wound on twin hole bead Philips grade 4B1

Code No. 4312 020 31525

Primary 4 turns of 2x.45mm. enamelled Cu. wire in parallel, tapped at centre and 2 x 1 turn from centre.

T_2 = 1:1.4 turns ratio.

Wound on a Philips 4C6 toroid 23x14x7mm.

Code No. 4322 020 91070

Primary 22 turns of 2x.45mm. enamelled Cu. wire centre tapped.

Secondary 16 turns of copper tape approx. 1.5mm. wide.

Primary wound on top of secondary winding and insulated by p.t.f.e. tape approximately .025mm. thick.

CH_1 2.5 turns on 6 hole bead grade 3B

Code No. 4312 020 31500

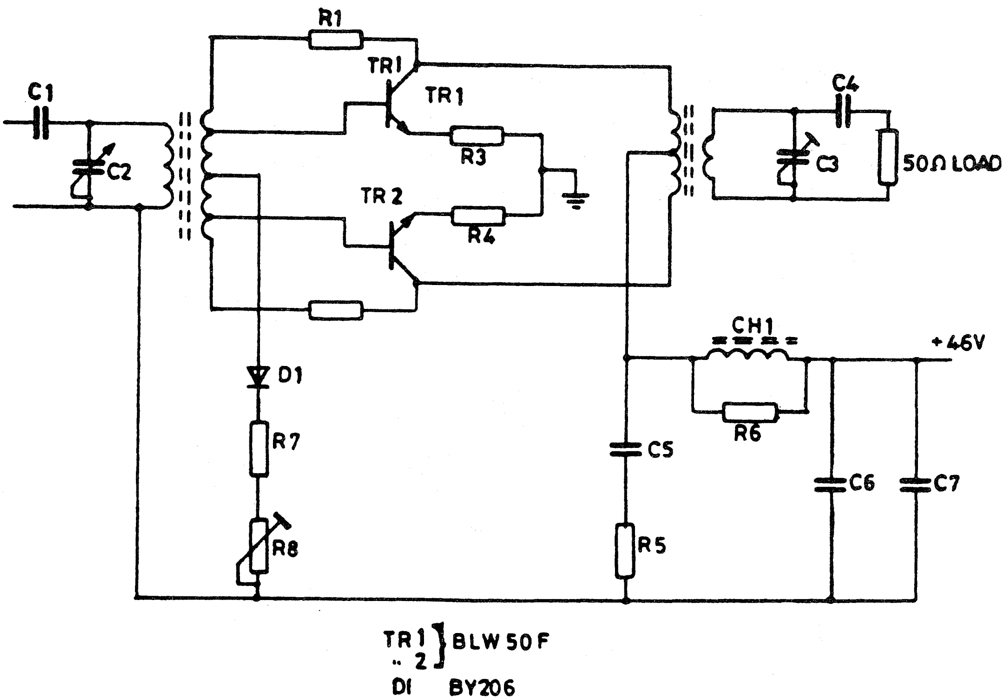

Fig. 1. Circuit Diagram

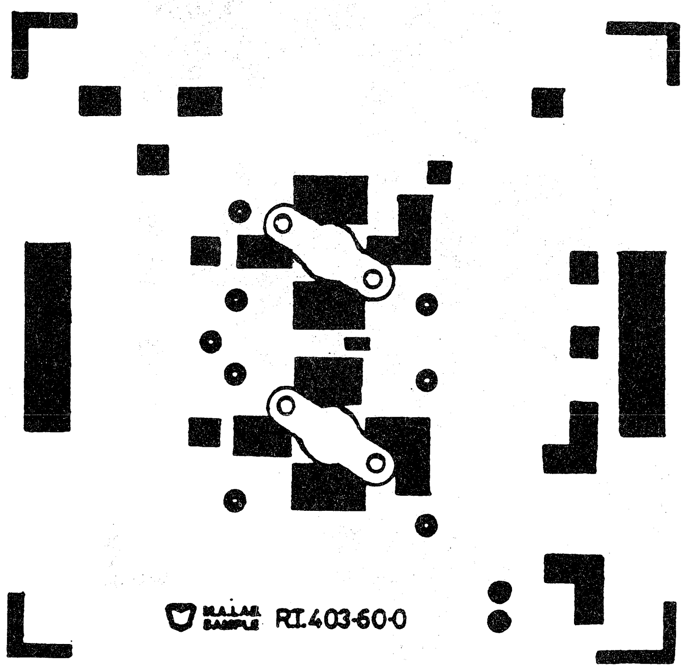

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Fig. 2. Circuit Board

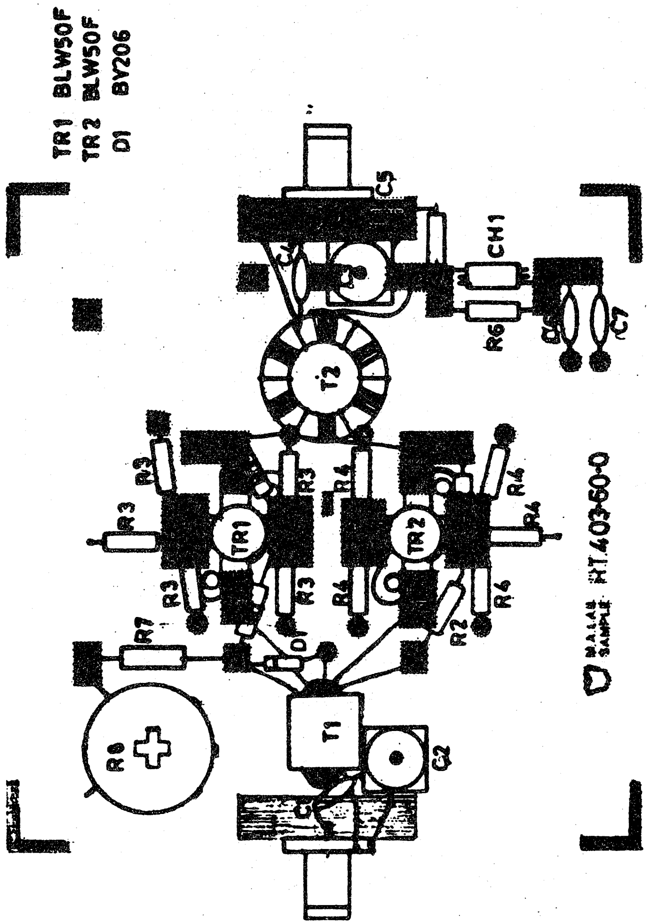

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Fig. 3. Component Layout

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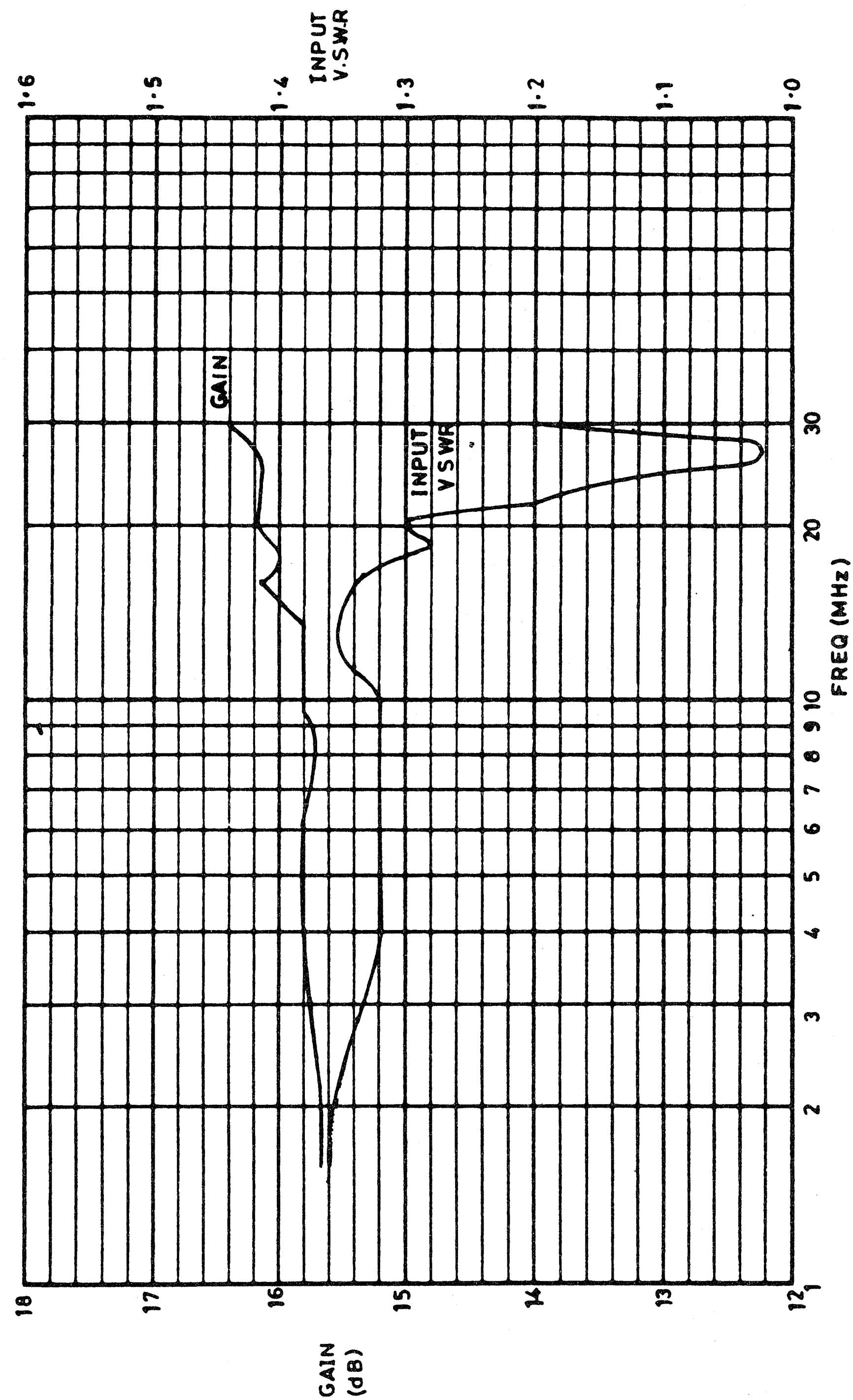

Fig. 4. BLW50F Linear Amplifier Gain and Input V.S.W.R vs Frequency

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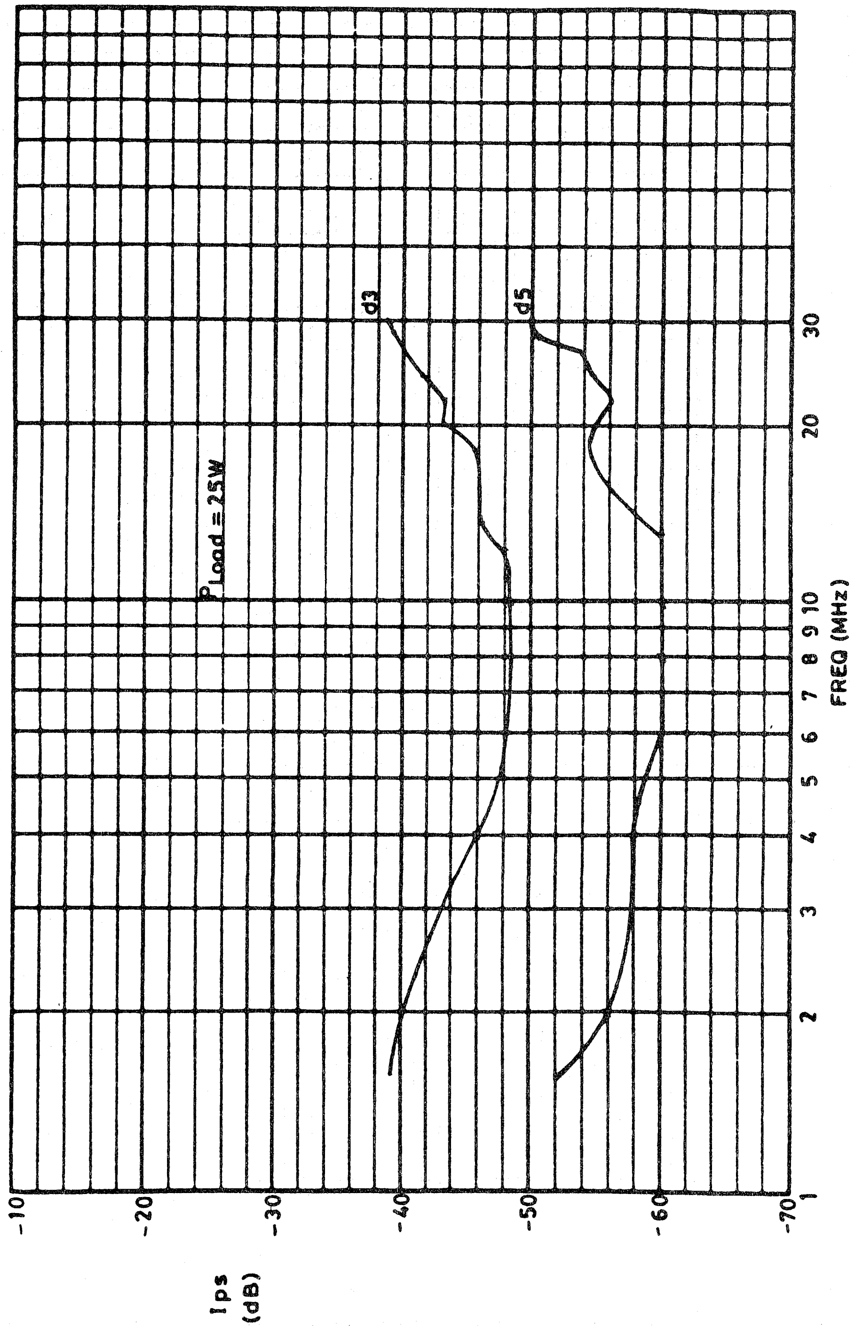

Fig. 5 . BLW50F Linear Amplifier-Intermodulation products vs. Frequency

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laboratory report

number : MC08004 **date** : 1980 jul 03

title : A Two Stage Wideband H.F.. Linear Applifier For 400 Watts P.E.P. Using BLW96 and BLW50F Transistors.

author : J. Ling

Mullard

laboratory report

from the applications laboratory:

Mullard application laboratory,
Mitcham, England.

12 AUG. 1980

number: MC08004	date: 1980 JUL 03
project:	pages: A1 ; R20 ;

title: A Two Stage Wideband H.F. Linear Amplifier For 400 Watts P.E.P. Using BLW96 And BLW50F Transistors.

author: J. LING approved: A.J. REES

ABSTRACT

A wideband linear h.f. amplifier is described which uses two BLW96 transistors in class AB, driven by two BLW50F transistors in class A. The amplifier operates from a 50 volt supply rail and will deliver up to 400 watts p.e.p. into a wideband 50 ohm resistive load with intermodulation products better than -26dB over the band 1.6MHz to 30MHz with overall gain over the band of 28dB ± 2dB.

The amplifier may be operated at 300 watts p.e.p., giving intermodulation products better than -30dB, including operation with the supply voltage reduced to 45 volts.

JL/AJR/MAT

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A TWO STAGE WIDEBAND H.F. LINEAR AMPLIFIER FOR 400 WATTS P.E.P. USING BLW96 AND BLW50F TRANSISTORS

1. INTRODUCTION

It has been shown [1] that two BLW96 transistors in a wideband class AB push-pull amplifier can give 400 watts p.e.p. under two-tone drive with intermodulation products <-26dB in the band 1.6 - 30MHz.

Also a suitable drive amplifier has been described [2] using two BLW50F transistors in class A at the same supply voltage (50V nominal). The intermodulation performance of the drive amplifier is <-40dB into a 50 ohm load.

The two amplifiers have been combined using direct inter-stage impedance transformation and the overall design is described in this report.

2. CIRCUIT DESCRIPTION

The circuit diagram of the complete amplifier is shown in Figure 1.

Design practice for the individual BLW96 and BLW50F push-pull linear amplifiers has been closely followed. The differences are :-

(i) Direct impedance matching between the driver stage output (100 ohms) and the p.a. input (5.5 ohms).

(ii) Replacement of adjustable capacitors by fixed values, because it is considered impractical to have the complication of adjustments in a two stage circuit.

The fixed capacitors are chosen to compensate for leakage reactance of critical transformers towards the higher frequency limit of the band, assuming good winding and mounting technique.

(iii) Omission of the balun and adjustable centre tap arrangement in the p.a. input circuits, which are unnecessary with balance provided by a push-pull driver.

3. CONSTRUCTIONAL DETAILS

Figures 2 and 3 show the printed circuit board and component lay-out.

Figure 4 shows the general arrangement of a water cooled copper heat sink on the underside of which the temperature compensated p.a. bias unit is mounted.

The parts list (Appendix) includes winding instructions and inductors and transformers.

4. AMPLIFIER PERFORMANCE

The separate performance of the basic driver and p.a. circuits with 50 ohm terminations are summarised below. We have :-

Bias

	Driver	P.A.
V_{CE}	44V	50V
I_C	2 x 1A	2 x 100mA (zero signal)

Single Tone
1.6 - 30MHz

P_L	25W	400W
Gain		
(mid band)	15.8dB	13.4dB
(min, max)	15.7dB, 16.4dB	13.4dB, 15.8dB
Input v.s.w.r.		
(mid band)	1.35:1	1.1:1
(max)	1.36:1	1.45:1

Two Tone;
1.6 - 30MHz

P_L (p.e.p.)	25W	400W
Efficiency		37.7% (min)
3rd order intermodulation		
(mid band)	-46dB	-28dB
(max)	-39dB	-27dB

4.1. Gain, V.S.W.R. And Intermodulation

The overall performance of the two stage amplifier is shown in Figures 5 - 9.

Figure 5 shows gain and input v.s.w.r. under single tone drive with 50V supply and P_{load} = 400 watts. Minimum gain and v.s.w.r. are seen to be 25dB and 2:1 respectively. Figures 6 - 8 show 3rd and 5th order intermodulation under two tone drive conditions with 50 volt supply and P_{load} 200, 300, 400 watts p.e.p.

It is seen that at 400 watts p.e.p., d_3 < -26dB and at 300 watts d_3 < -30dB.

In addition, figure 9 shows intermodulation performance at 300 watts p.e.p., but with 45V supply. It is seen that d_3 is still < -30dB.

4.2. Harmonic Content

The amplifier was driven to 400 watts c.w. and the amplitude of the harmonics measured relative to the fundamental signal in the wideband load.

f (MHz)	f_2 (dB)	f_3 (dB)	f_4 (dB)	f_5 (dB)	f_6 (dB)	f_7 (dB)	f_8 (dB)	f_9 (dB)	f_{10} (dB)
1.6	-46	-19	-56	-34	-48	-48	-50	-45	-59
3.5	-45	-19	-50	-33	-58	-44	-56	-45	-60
7	-54	-18	-50	-29	-48	-40			
10	-48	-17	-45	-32	-55	-50			
14	-43	-16	-50	-44	-	-			
20	-34	-25	-	-	-	-			
28	-40	-45	-	-	-	-			

5. CONCLUSIONS

A wideband linear h.f. amplifier has been designed using BLW96 output and BLW50F driver transistors.

The overall amplifier gain is in the range 25-29dB over the band 1.6 to 30MHz and the input v.s.w.r. is < 2:1.

When operated at 400 watts p.e.p. (two-tone) intermodulation products are < -26dB.

Intermodulation product < -30dB may be obtained at 300 watts p.e.p. even with the supply rail reduced to 45 volts.

6. REFERENCES

(1) J. Ling. A single stage wideband (1.6 - 30MHz) linear amplifier for 400 watts p.e.p. using BLW96 transistors.

M.A.L. Report MC08002.

(2) J. Ling A single stage wideband (1.6 - 30MHz) linear amplifier for 25 watts p.e.p. using BLW50F transistors in class A.

M.A.L. Report MC08003.

A P P E N D I X

PARTS LIST

Resistors

				Code No.
R_1 R_2	=	5 x 12Ω ± 5% in parallel	Philips type	CR25, 2322 211 13129
R_3 R_4	=	2 x 200Ω ± 5% in series	" "	PR52, 2322 192 32001
R_5	=	15Ω ± 5%	" "	CR25, 2322 211 13159
R_6 R_{18}	=	3.3Ω adjustable	" "	TPW22, 2322 011 02338
R_7 R_{14} R_{15}	=	22Ω ± 5%	" "	CR25 2322 211 13229
R_8	=	6.8Ω ± 5%	" "	CR25, 2322 211 13688
R_9	=	1.8Ω ± 10%	" "	AC10, 2322 329 10188
R_{10} R_{11}	=	18Ω ± 5%	" "	PR37, 2322 212 31209
R_{12} R_{13}	=	2 x 12Ω ± 5% 2 x 15Ω ± 5% } in parallel	" "	PR37, 2322 212 31209 PR37, 2322 212 31509
R_{16}	=	1.5kΩ ± 5%	" "	PR37, 2322 212 31502
R_{17}	=	3 x 180Ω ± 5% in parallel	" "	EH15, 2322 330 03181
R_{19}	=	22Ω ± 5%	" "	CR37, 2322 212 13229

Capacitors

					Code No.
C_1 C_{23} C_{24}	=	10nF ± 20% polyester.	Philips	type	352, 2222 352 44103
C_2	=	39pF ± 2% ceramic.	"	"	632, 2222 632 10399
C_3	=	27pF ± 2% ceramic.	"	"	632, 2222 632 10279
C_4	=	680pF ± 2% polystyrene.	"	"	426, 2222 426 6801
C_5	=	22nF ± 20% polyester.	"	"	352, 2222 352 44223
C_6	=	2 x 47nF ± 20% in parallel polyester.	"	"	352, 2222 352 42473
C_7 C_8	=	2 x 470pF ± 2% 1 x 390pF ± 2% } in parallel polyester	"	"	426, 2222 426 4701 426, 2222 426 3901
C_9 C_{10}	=	2 x 1000pF ± 2% 1 x 820pF ± 2% } in parallel polyester	"	"	426, 2222 426 1002 426, 2222 426 8201
C_{11}	=	100nF ± 20% polyester	"	"	352, 2222 352 44104
C_{12} C_{13}	=	2 x 47pF ± 2% 2 x 56pF ± 2% } in parallel ceramic	" "	" "	632, 2222 632 34479 632, 2222 632 34569
C_{14} C_{15}	=	5 x 10nF ± 20% in parallel polyester.	"	"	352, 2222 352 44103
C_{16} C_{17}	=	3 x 100nF ± 20% in parallel polyester.	"	"	352, 2222 352 54104
C_{18}	=	60pF trimmer	"	"	809, 2222 809 08003
C_{19} C_{20}	=	3.3µF ± 10% polyester.	"	"	344, 2222 344 21335
C_{21} C_{22}	=	220µF 4V electrolytic	"	"	016, 2222 016 2221
C_{25}	=	220µF 10V electrolytic	"	"	016, 2222 016 4221

Inductors

Ch_1 CH_2 CH_3	2.5 turns through 6 hole ferrite bead grade 3B code no. 4312 020 31500.
Ch_4 Ch_5	3 parallel loops through 6 hole ferrite bead grade 3B code no. 4312 020 31500.
L_1 L_2	13.9nH - see diagram on Figure 1.
L_3 L_4	21nH. - see diagram on Figure 1.

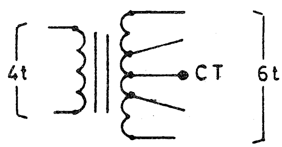
Transformers

T_1 1:1.5 turns ratio.

Wound on twin hole bead Philips grade 4B1. Code no. 4312 020 31525.

Primary: 4 turns of 2 x .45mm enamelled Cu wire in parallel.

Secondary: 6 turns of 2 x .45mm enamelled Cu wire in parallel tapped at centre and at 2 x 1 turns from centre.



Typical primary reactance at f = 1.6MHz = (secondary o/c) = j160.

Typical leakage reactance at f = 30MHz = (secondary s/c) = j25.

T_2 4.5:1 turns ratio.

Consists of two transformers with primary and secondary windings connected in parallel, each wound on twin hole bead Philips grade 4B1, code no. 4312 020 31500.

Primary winding 9 turns 0.45mm enamelled Cu wire centre tapped.

Secondary 2 turns of 2 x 0.45mm enamelled Cu wire in parallel. Typical primary reactance of combination at f = 1.6MHz (secondary o/c) = j400.

Typical leakage reactance of combination at f = 30MHz (secondary s/c) = j90.

T_3 The centre tapped choke

Wound on a 50mm length of 4A10 aerial rod (or equivalent), code no. 4311 020 55390.

4 turns of twisted enamelled Cu wire 1.0mm.

Typical total reactance at f = 1.6MHz = j40.

T_4 2.33:1 turns ratio.

Consists of two transformers with primary windings and secondary windings connected in parallel, each wound on 4C6 toroids 36 x 23 x 15mm, code no. 4322 020 91090.

Primary winding 6 turns of Cu tape 8mm wide.

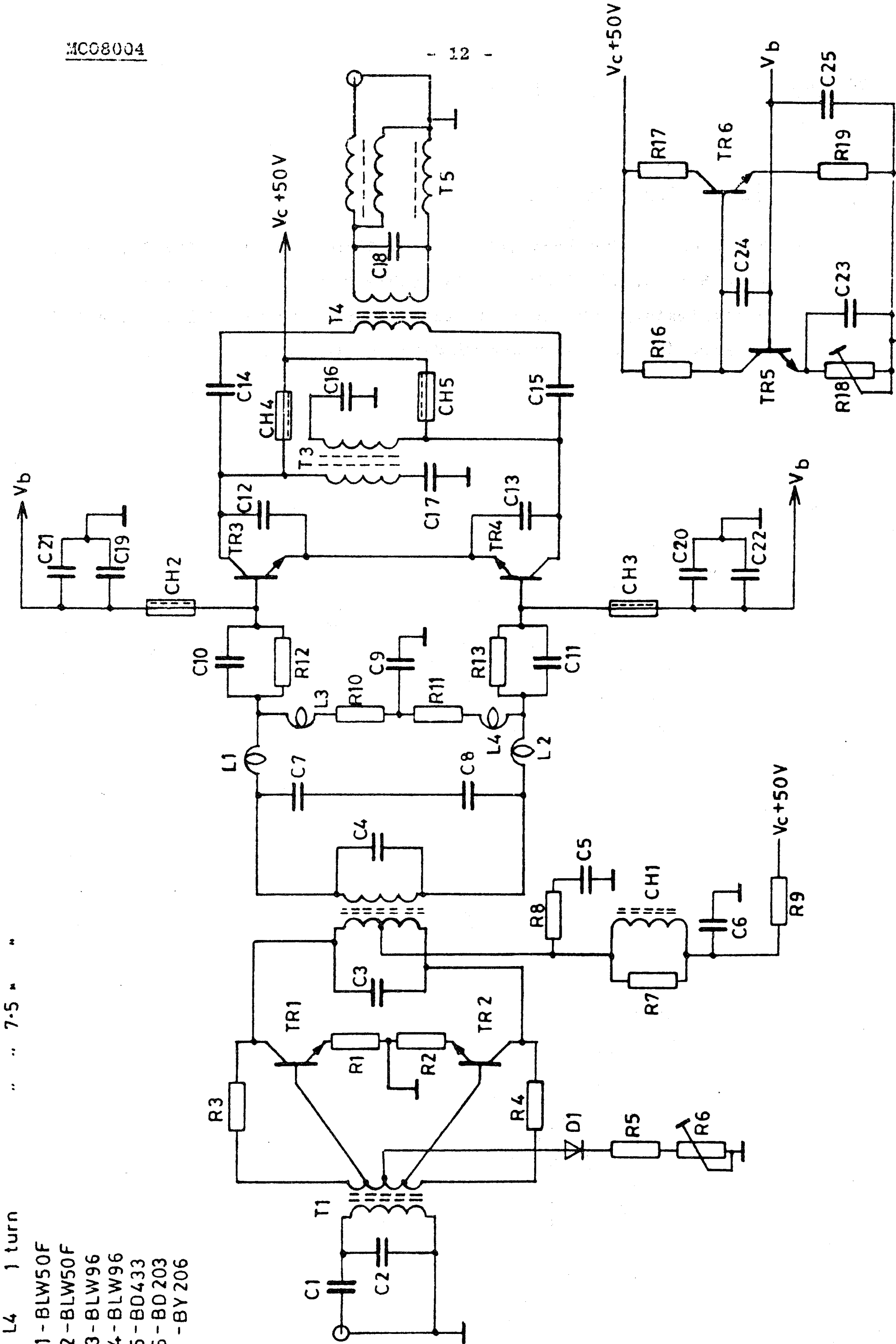
Secondary winding 14 turns of 4 x 0.5mm enamelled Cu wire in parallel. Windings are separated by p.t.f.e. tape approximately 0.25mm thick.

Typical primary reactance of combination at f = 1.6MHz (primary o/c) = j200.

Typical leakage reactance of combination at f = 30MHz (primary s/c) = j50.

T_5 Output balun transformer

Wound on two 4C6 toroids 36 x 23 x 15mm, code no. 4322 020 91090, with 8 turns of 50 ohm coaxial cable having p.t.f.e. insulation and approximately 4mm external diameter and 8 turns of 1mm enamelled cu wire for the balancing winding - see diagram.



L1, L2 1 turn 1·3 mm cu wire 5 mm dia.
L3, L4 1 turn " " 7·5 " "
TR1 - BLW50F
TR2 - BLW50F
TR3 - BLW96
TR4 - BLW96
TR5 - BD433
TR6 - BD203
D1 - BY206

Fig. 1. Circuit Diagram

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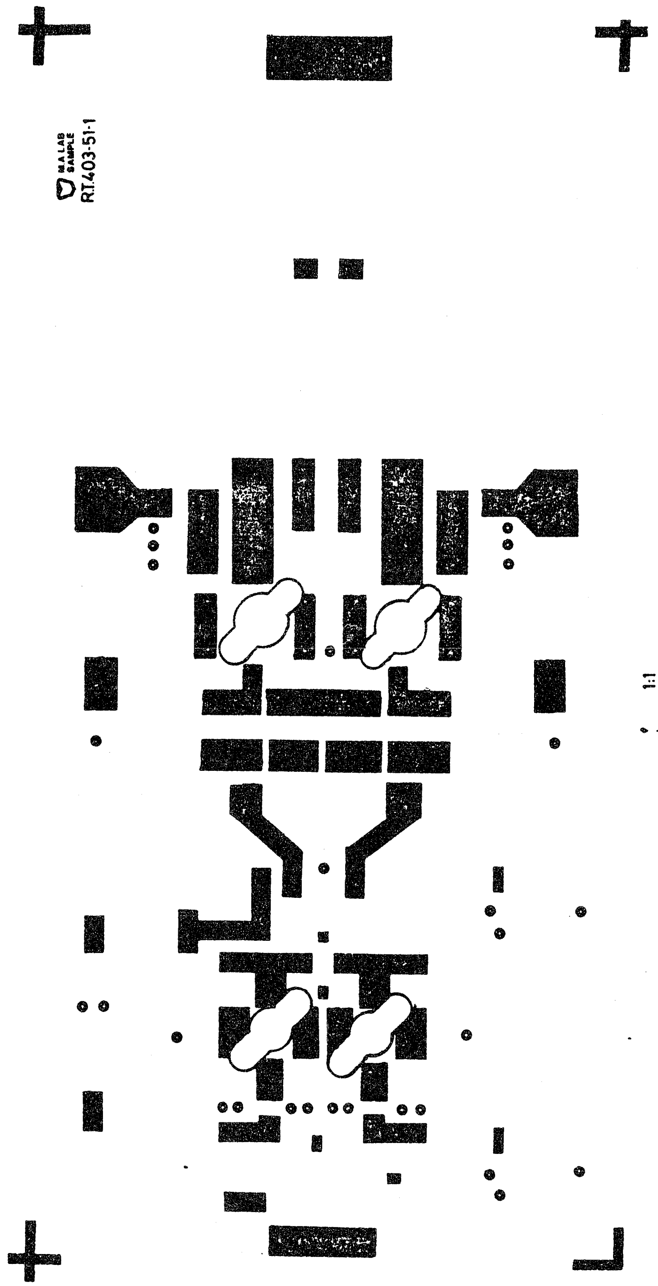

Fig. 2. Printed Circuit Board

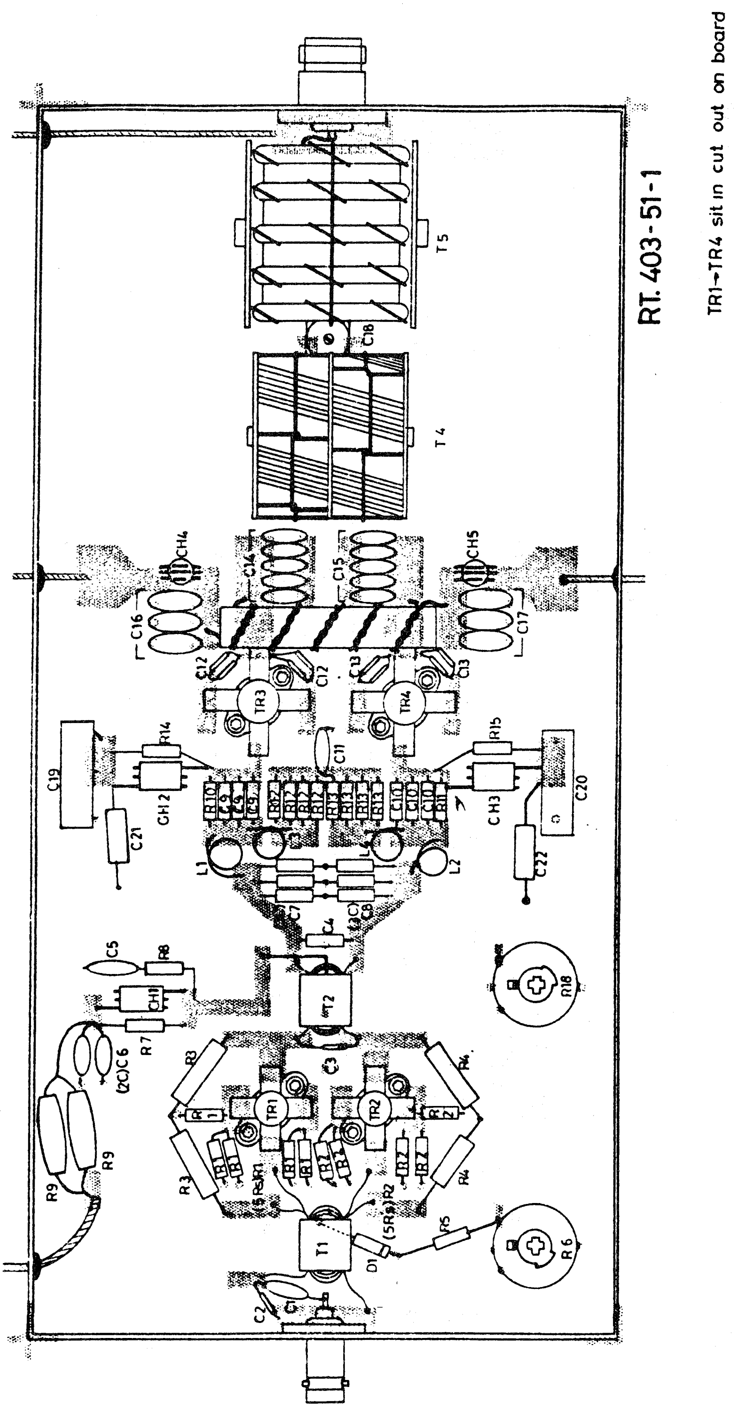

Fig. 3. Component Layout

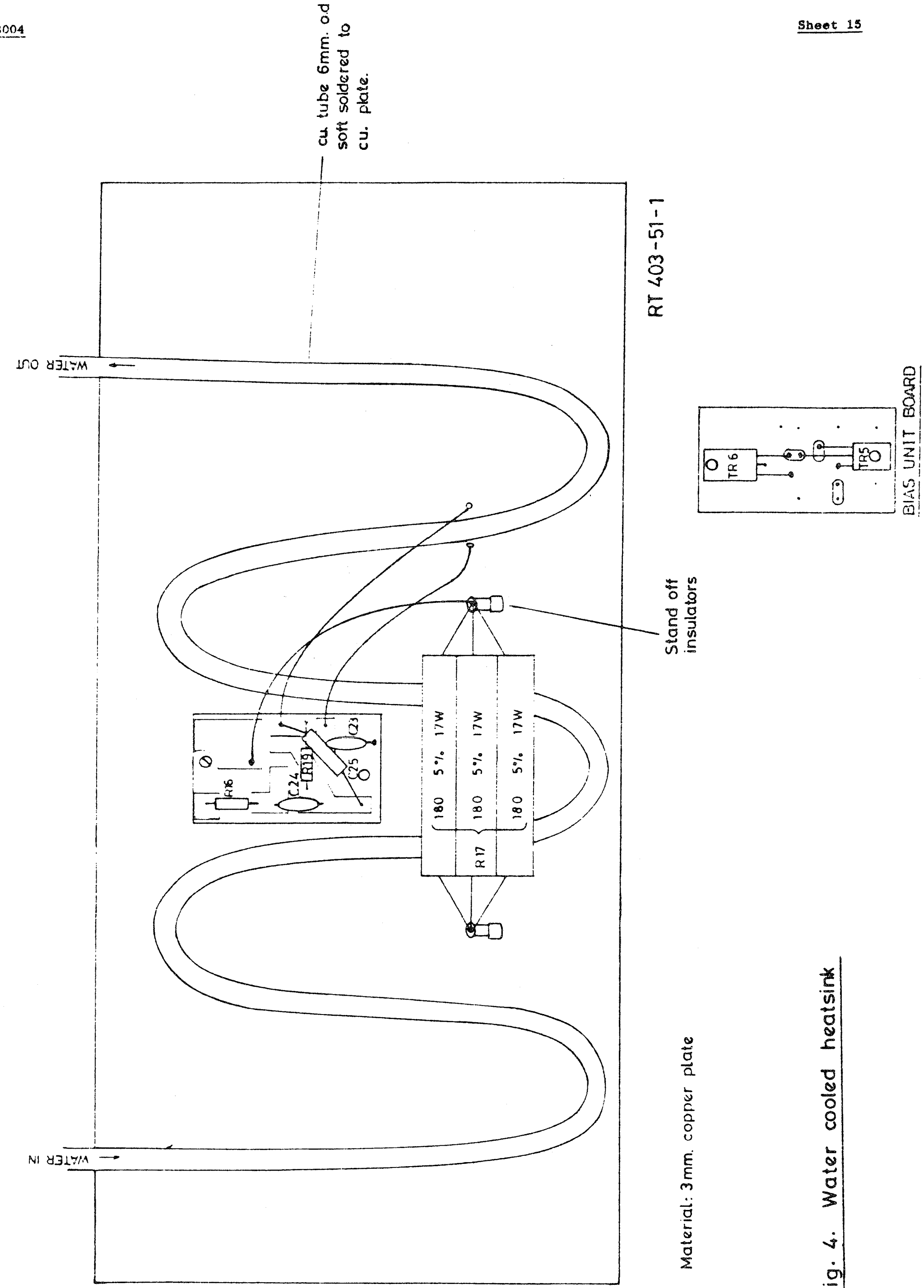

Fig. 4. Water cooled heatsink

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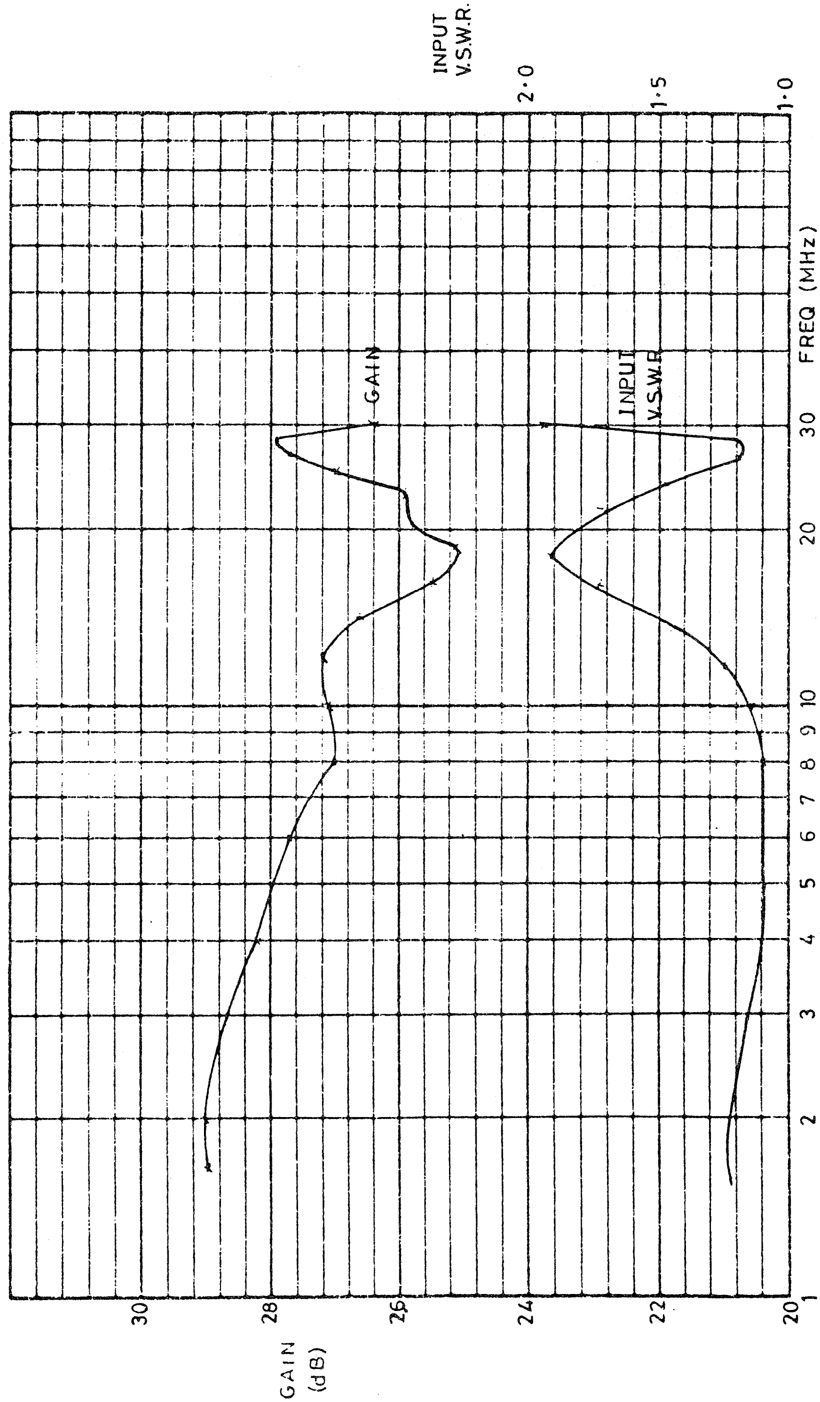

Fig.5 . Amplifier Gain & V.S.W.R

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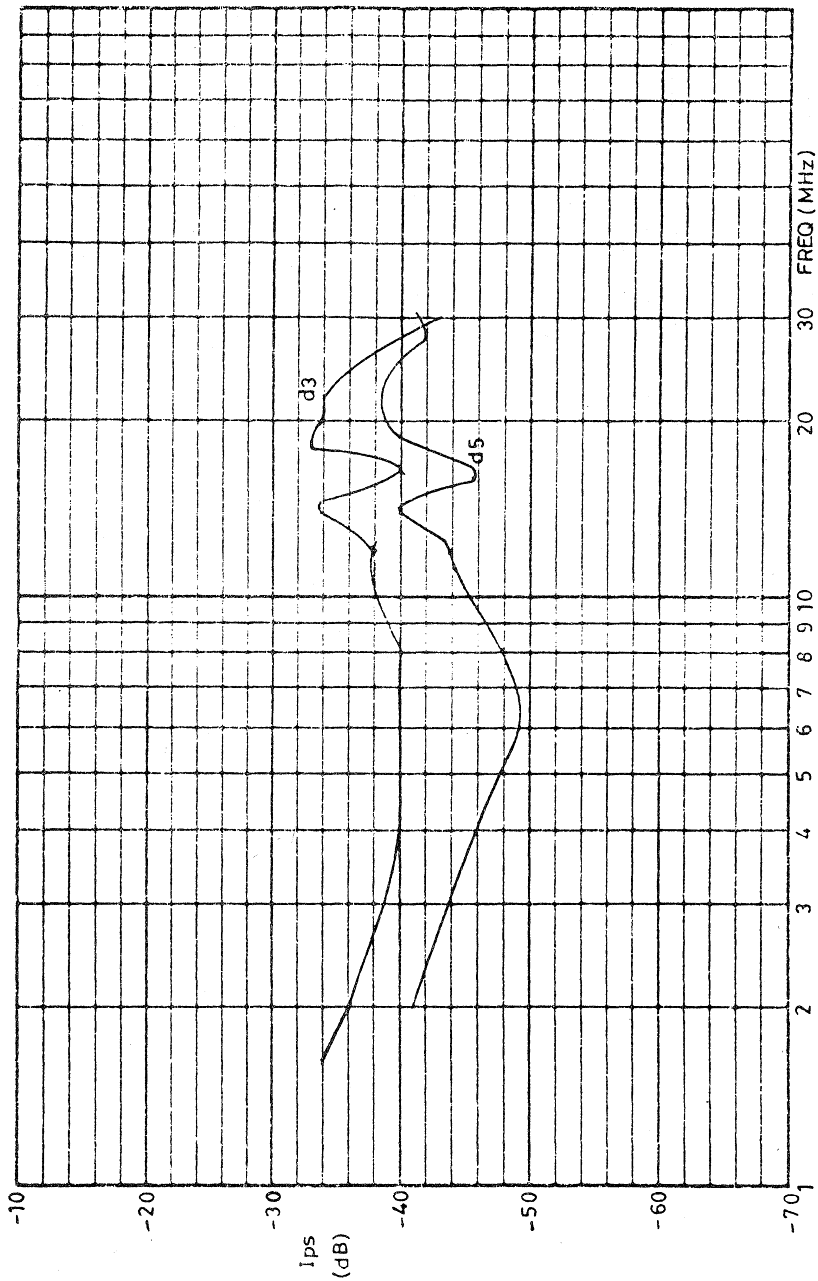

Fig. 6. Intermodulation Performance (50V, 200W p.e.p)

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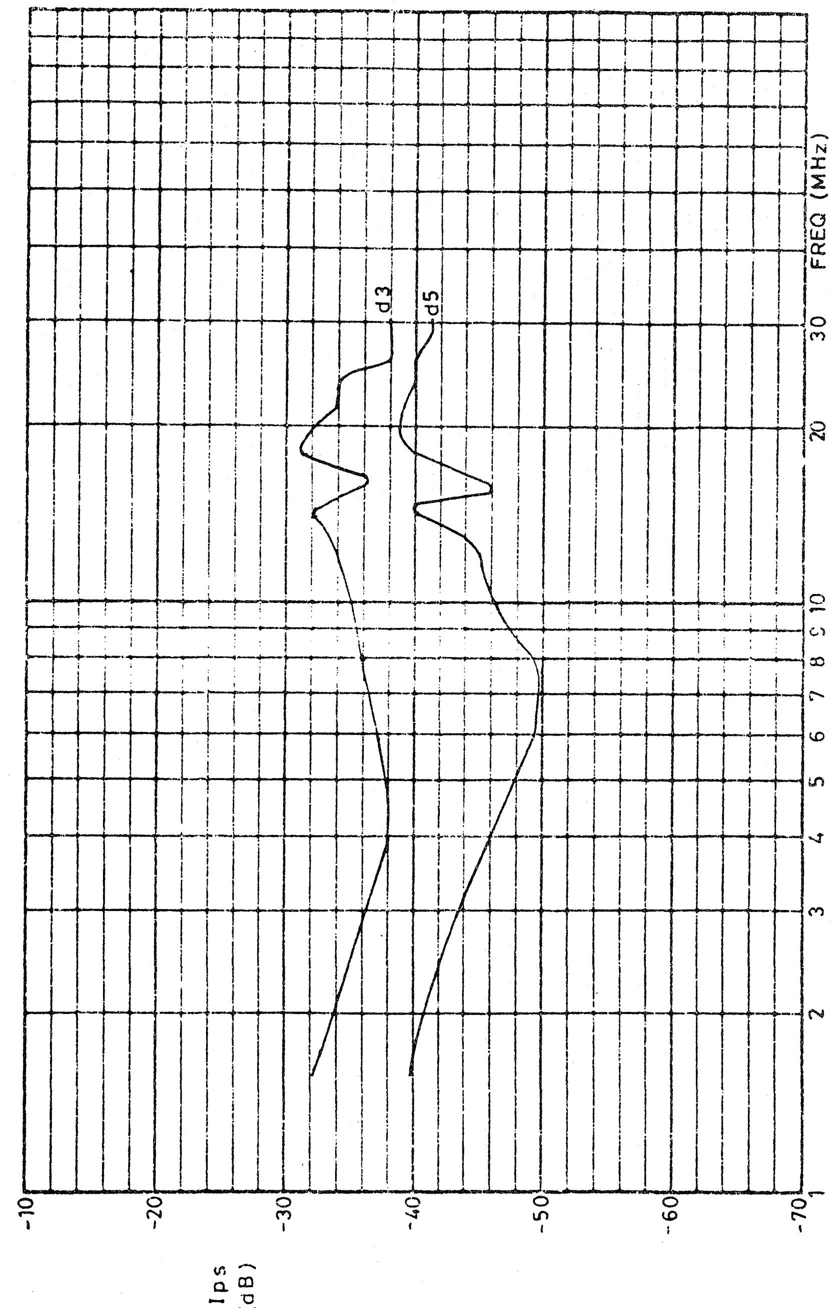

Fig. 7. Intermodulation Performance 50V 300W p.e.p

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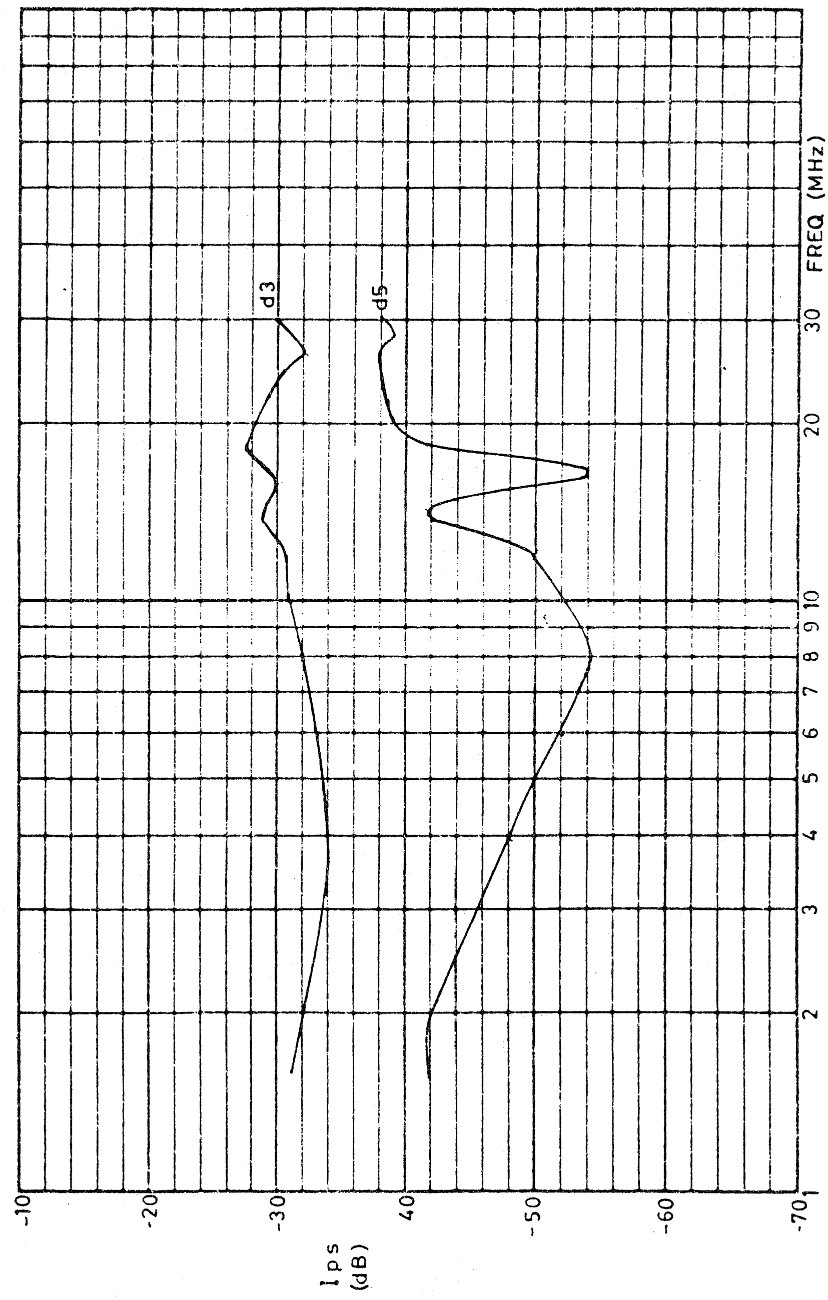

Fig. 8. Intermodulation Performance (50V. 400 p.e.p)

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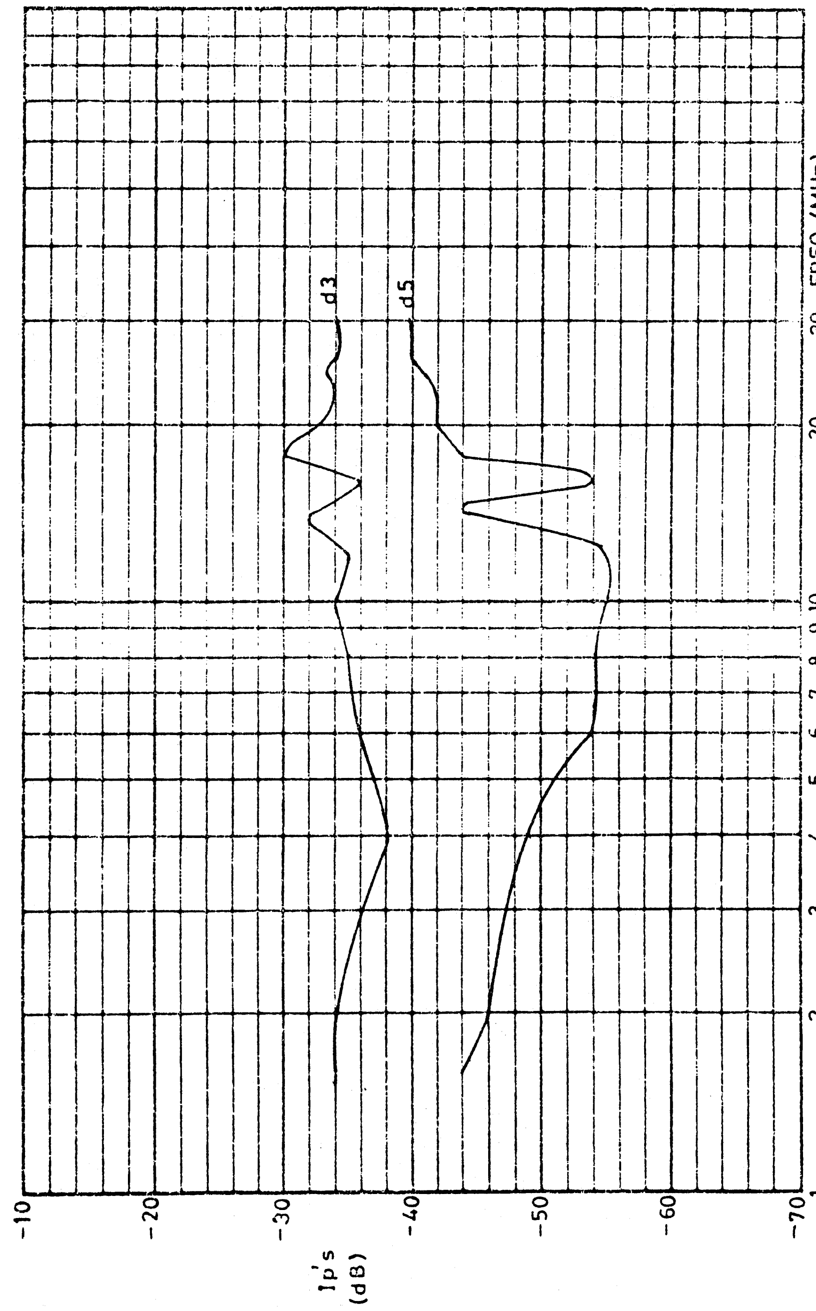

Fig. 9. Intermodulation Performance (45V, 300W pep)

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N.V. PHILIPS SEMICONDUCTORS APPLICATION LABORATORY
NIJMEGEN – THE NETHERLANDS

REPORT No: NCO 8203	AUTHOR: A.H.Hilbers
PROJECT No: 5579	DATE: 1982-06-08

TITLE

DESIGN OF A 1kW GENERATOR FOR INDUSTRIAL HEATING APPLICATION

ABSTRACT

In this report a description is given of a simplified amplifier module with 2 transistors BLW 96 producing a maximum output power of 400 W at 13.56 MHz and 27.12 MHz.
The maximum supply voltage is 50 Volt.
In addition some system aspects are being considered, e.g. combination to higher power levels, line-ups for drivers and power control to obtain fail-safe operation during load mismatch.

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Advice Patents Dept. d.d: 24.08.'82	AV	GV		B		BL	
Decision MAMO d.d: 20-07-'82	AV	GV	EI	B		BL	
DATE: 28 JULI 1982	MAMO:						

DESIGN OF A 1KW GENERATOR FOR INDUSTRIAL HEATING APPLICATION

1. INTRODUCTION

The transistor BLW 96 has primarily been designed for application in HF-SSB transmitters in the frequency range 1.5-30MHz.
At a supply voltage of 50V the device is able to produce an output power of 200W.
In application report MCO 8002 (Ref.1) a wideband amplifier has been described for the frequency range 1.5-30MHz.
It contains 2 transistors BLW 96 operating in class-AB and its maximum output power is 400W.
For industrial heating purposes this amplifier is unnecessarily complicated because no linearity requirements exist and operation is only required at either 13.56 MHz or 27.12 MHz.
It is the main intention of this report to indicate the possible simplifications of the above mentioned amplifier.
Additionally some other points will be considered like combination to higher power levels, line-ups for drivers and power control for obtaining fail-safe operation during load mismatch.

2. SYSTEM CONSIDERATIONS

Most questions on industrial heating applications range from a few hundred Watts to some Kilowatts.
Fig.1 shows an example of a line-up for an output power of appr. 1 KW.
The final stage consists of 4 modules each with 2 transistors BLW 96, operating in class-B. This module will be described in more detail in the next section.
The combination of the modules can be done with hybrid couplers as described in report MCO 7404 (Ref.2) and COM 74133 (Ref.3).
The output impedance of the combined power amplifier is 50 ohms.
It is followed by a directional coupler needed for control of the output power and a tuning unit for matching the 50 ohms output impedance of the power amplifier to the actual load impedance.

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The maximum output power of the combination is appr. 1300W at a supply voltage of 50V. This power level can be maintained up to a load VSWR of 2.
At 27.12 MHz the combined output stage requires a drive of appr. 70W which can be delivered by an amplifier containing 2 transistors BLW 50F and operating in class-B push-pull from the same supply voltage as the final stage. The pre-driver and X-tal controlled oscillator need a lower supply voltage, e.g. 13V.
At 13.56 MHz the output stage requires a lower drive power, i.e. appr.20W. This can for instance be achieved by a reduction of the supply voltage of the pre-driver and X-tal oscillator stages. Alternatively an attenuator can be connected between the pre-driver and driver stages.
An important point to be considered is that of heavy load mismatch. The best way to cope with this phenomenon is a simultaneous reduction of the supply voltages of final stage and driver.
This action must start at a load VSWR of 2. A suitable end point is a supply voltage of 40V at a load VSWR of 50. The forward output power has then be reduced to appr.800W. A very effective method of protection has been described by Mr.K.Ruf (see Ref.4). He uses a dual directional coupler and his control system is arranged in such a way that the sum of the forward and reflected voltage is kept constant. Consequently the forward output power at severe load mismatch will be reduced to one fourth of the original value, i.e. to appr. 300W. The required supply voltage reduction is then appr. 50%.

3. FINAL STAGE AMPLIFIER MODULE

Compared with the SSB amplifier module as described in Ref.1 the following simplifications can be made:

A. A class-AB bias unit is not required.

B. In the output network the collector choke and coupling capacitors can be omitted.

C. In the input network a number of components required for obtaining a flat power gain can be omitted.

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The final result is shown in Fig.2 and the parts list.
For the construction of the impedance transformers one is referred to report ECO 7308 (Ref.5).
The maximum output power per module is 400W at a supply voltage of 50V. The required drive power is appr.15W at 27.12 MHz.

REFERENCES

1. J.Ling, "A Single stage wideband (1.6MHz to 30MHz) Linear Amplifier for 400W P.E.P. Using BLW 96 Transistors, Application report MCO 8002.

2. J.Ling, "A 1.0KW P.E.P. Linear Power Amplifier from f=1.6MHz to 30MHz Using BLX 15 Transistors", Application report MCO 7404.

3. J.Ling,"Alternative Combining Unit for 1KW Wideband Amplifier Using BLX 15 Transistors", Technical Note COM 74133 [1)].

4. K.Ruf,"Leistungsregelung und Leistungsbilanz Transistorisierter Sendeverstärker", Nachrichten Elektronik 12,1979,pp 400-402.

5. M.J.Köppen,"A single Stage Wideband (1.6-28MHz) Linear Power Amplifier for 300W P.E.P. Using 2x BLX 15, Application report ECO 7308.

1) Can be ordered from Mullard, London or Philips, Nijmegen.

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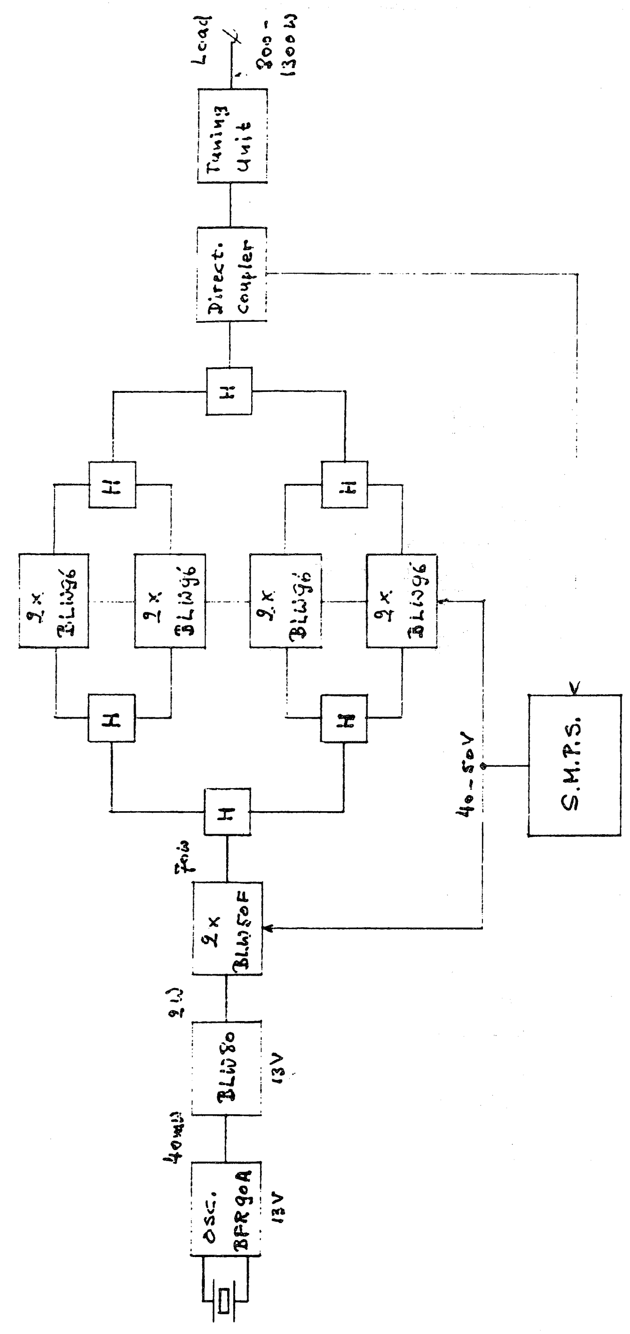

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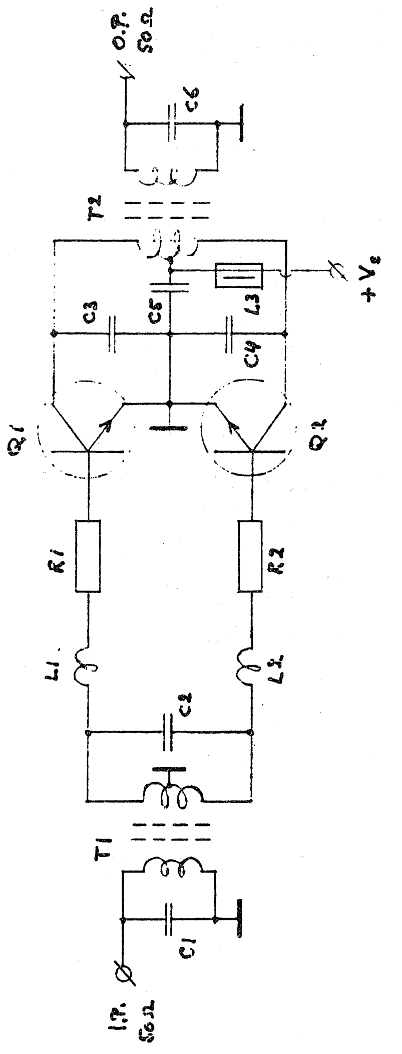
Electronic components and materials



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PARTS LIST

Q1=Q2= BLW 96

R1=R2= Parallel connection of 5 resistors of 2.2. ohms, PR 37 type.

C1= 270pF, ceramic 500V.

C2= parallel connection of 2x680pF and 820pF, polystyrene type.

C3=C4= parellel connection of 150pF and 120pF,ceramic 500V.

C5= 100nF, polyester.

C6=parallel connection of 2x22pF and 27pF, ceramic 500V.

L1=L2= 14.3nH;0.5 turn of 1 mm copper wire , D_{int}=6 mm, leads 2x7 mm.

L3= Fxc choke grade 3B, cat.nr.4312 020 31500, wound with 6 wires in parallel.

T1= 1:3 transformer on Fxc toroid grade 4c6,
dimensions $14x9x5mm^3$,cat.nr.4322 020 91020
primary: 21 turns of 0.5mm enamelled copper wire
secondary: 7 turns of 75 µm copper foil,width= 3 mm

T2= 2:1 transformer on Fxc toroid grade 4c6,
dimensions $36x23x15mm^3$, cat.nr.4322 020 91090
primary: 6 turns of 75 µm copper foil, width = 10mm
secondary: 12 turns of 2x0.7mm enamelled copper wire.

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Report no : RNR-1-259-1987-AS / NCO 8703
Author : G.Lukkassen
Date : 1987-04-28

A WIDEBAND LINEAR POWER AMPLIFIER (1.6-28MHz) FOR 300W PEP WITH 2 MOS TRANSISTORS BLF177

SUMMARY

This report gives a description of a wideband push-pull amplifier for the frequency range 1.6-28MHz.

The amplifier has been designed around 2 MOS transistors BLF177 which operate in class-AB at V_{DS}=50V and I_{DQ}=0.5A/transistor.

The main properties at P_o = 300W are:

Powergain	:22-23dB
Efficiency	:52.5-61%
Return losses input	:$\leqslant$-15.5dB
2nd harmonics	:$\leqslant$-25dB
3rd harmonics	:$\leqslant$-16dB
IMD at P_o=300W PEP	:$\leqslant$-33dB

NEDERLANDSE PHILIPS BEDRIJVEN BV - PRODUCT DIVISION ELCOMA - THE NETHERLANDS

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1. INTRODUCTION

The BLF177 is an RF Power MOS transistor for the HF and VHF range in a 4 leads flange SOT121 encapsulation. For the frequency range 1.6-28MHz a wideband push-pull power amplifier has been developed with 2xBLF177 having an output power of 300W PEP at an intermodulation distortion level below -30dB.
The transistors operate in class-AB at V_{DS}=50V and a quiescent current of 0.5A each.

2. DESIGN OF THE AMPLIFIER

2.1. General

The schematic set-up is given in Fig.1.

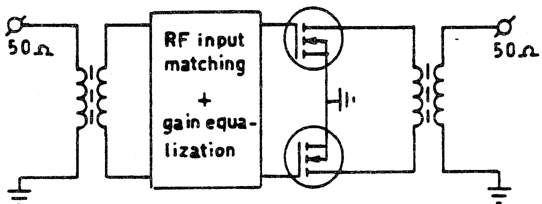


Fig. 1 Schematic set-up 2x BLF177 amplifier.

The two balance to unbalance transformers are applied to split the single ended input into 2 out of phase driving ports and to add the 2 out of phase output ports into one single ended output. The transformers have an impedance transformation ratio of 4:1 and match the low-ohmic in-and output impedance of the transistors to the 50Ω system impedance. At the input a special circuit takes care of a good input matching and a flat powergain over the whole bandwith.

2.2. Output circuit

2.2.1. Load impedance

The output impedance of each transistor can be represented as a combination of the output capacitance C_{oss} and the optimum load resistance. Because of the larger drain voltage swing the effective output capacitance C_o is appr. 15% higher than the value of C_{oss}. So C_o=1.15*190≈220pF. The optimum load resistance for class-AB can be determined with formula:

$$R_L=(0.85*V_{DS})^2/(2*P_o).$$

For V_{DS}=50V and P_o=150W we get $R_L \approx 6\Omega$. To keep the transformer simple a transformation ratio of 4 or 9 is preferable. A ratio of 4 gives a load impedance of 50/4=12.5Ω → 6.25Ω for each transistor. This is very near to the optimum load resistance.

2.2.2. Output transformer

The output transformer has to transform the 50Ω asymmetrical impedance to the 2x6.25=12.5Ω symmetrical load impedance. The reactance (ωL) of the shunting inductance at 1.6MHz has been chosen at 4 times 50Ω=200Ω. So the inductance is 20μH. The transformer has been wound on a ferrite toroid of 4C6 material. Dimensions: 36x23x15mm (Dxdxh) which gives a volume (A.l)=8.97E-6m^3.
Because the power handling of one toroid is critical two transformers in parallel with an inductance of 40μH each have been chosen.

n_{sec}=SQR((L.l)/($\mu_o \cdot \mu_r$.A))

n_{sec}=SQR((40E-6*9.2E-2)/(4πE-7*120*97.6E-6))

=15.8 turns.
So n_{pr}=8 turns and n_{sec}=16 turns.

For each transformer V_{max} depends on the power over 100Ω.
V_{max}=SQR(2.P_o.R_L)=SQR(2*150*100)=173.2V.

B_{max} depends on the parallel loss resistance at 1.6MHz; for a power loss of 1%:
B_{max}=1.3E-2T.
The volume A.l needed per core is:

A.l=(V_{max}/(ω.B_{max}))^2($\mu_o \cdot \mu_r$)/L.

A.l=(173.2/(2π*1.6E+6*0.013))^2(4πE-7*120)/

40E-6=6.62E-6m^3.
Each of the toroids has a volume of 8.97E-6m^3. Fig.2 on page 16 shows one of the two parallel connected output transformers. On each toroid the primary winding has 8 turns of copperfoil (width 5mm and thickness 0.05mm). The secondary winding has 16 turns of 2 enamelled copper wires (0.6mm) in parallel.

So each primary turn has been covered with 2 secondary turns which means 4 wires of 0.6mm. Both windings are isolated with PTFE-foil (thickness 0.1mm). To reduce the stray-inductance the transformer has been wound as follows:

a) The primary has been wound evenly around the periphery of the toroid.
b) With the secondary the same has been done with the first 8 turns; the second part of 8 turns has been wound in between the first part. So the secondary has been wound twice around the core.

The measured secondary inductance of each transformer is 38μH and L_{str}=300nH.
With the aid of a network analyser the parallel combination of these 2 transformers has been corrected. For the higher frequencies at the low-ohmic side a parallel capacitor of 240pF and for the lower frequencies at the high-ohmic side a series capacitor of 10nF give return losses below -21dB over the whole frequency range (see Fig.3).

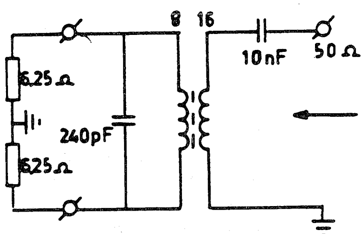


Fig.3 Output transformer with correction.

Replacing the transistors by resistors of 6.25Ω the return losses can be measured at the 50Ω side. Fig.4 on page 17 gives the return losses of the parallel combination of the two transformers before and after the correction.

2.2.3. The tapped choke

The chokes in the drain circuits are wound around a common ferrite rod of 4B1 material. Dimensions: 50x10mm (lxd). Fig.5 gives a schematic electrical circuit of the output.

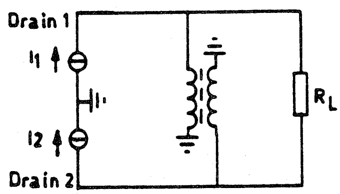


Fig.5 Output with the tapped choke.

Between both drains the impedance for the even harmonics depends on the coupling factor between both windings. If the coupling factor amounts to 1 both drains will be short circuited for the even harmonics.
Because the voltage over one winding is equal to half of the voltage between both drains, the total inductance between both drains is 4 times the inductance of one winding.
The reactance of the shunting inductance at 1.6MHz has been chosen at 4 times 12.5Ω=50Ω.
So the inductance between both drains is 5μH, this means for one winding an inductance of 1.25μH. According to the Philips Data Handbook "Components and Materials", C5,1986,page 320 the effective permeability of a rod with l/d=5 and μ_r=250 is appr. 20.
The number of turns can be calculated with:

$$n=SQR(L.l/(\mu_o.\mu_r.A)).$$

$n=SQR(1.25E-6*50E-3/(4\pi E-7*20*\frac{1}{4}\pi(10E-3)^2)=5.6$ turns. In practice 6 twisted turns of the primary and secondary windings have been wound around the rod. Fig.6 on page 16 shows the tapped choke. To increase the coupling factor each winding consists of 2 enamelled copper wires (0.8mm) in parallel. The measured inductance is 1.275μH.

2.2.4. Tuning of the output circuit

For an optimum alignment of the output circuit the 2 transistors have been replaced by dummies consisting of the parallel connection of a resistance and a capacitance. The resistance is equal to the optimum load resistance and the capacitance to the output capacitance (see section 2.2.1.).
Tuning of the output circuit has been carried out by measuring the return losses at the output with a network analyser under swept conditions (see Fig.7).

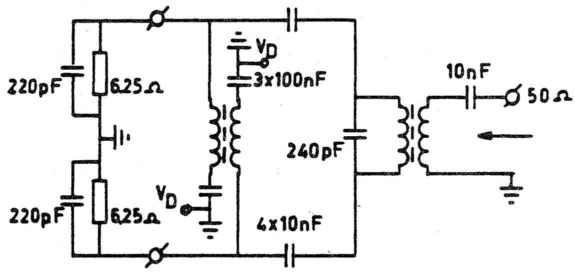


Fig.7 Output circuit before tuning.

The measured return losses should be as low as possible by changing the correction capacitors. Fig.8 on page 17 shows the return losses of the output before and after tuning. For optimum results the capacitance across the primary winding of the output transformer has been reduced from 240pF to 150pF and the low frequency correction capacitor of 10nF at the output has been changed to an inductance of 100nH. The last change can be explaned as follows:

a) The low frequency compensation is taken over by the coupling capacitors between the drain choke and the impedance transformer.
b) The function of the transformer is not only impedance matching but also transfer from balanced to unbalanced. The latter makes that the interwinding capacitance has more influence. This is so much that a series inductance at the output is needed for high frequency compensation.

2.3. Input circuit

2.3.1. Input impedance

The input impedance and gain of the transistor can be determined with the aid of a computer model of the BLF177. Fig.9 shows the calculated gain and impedances for the frequency range 1.6-28MHz.

```
Gf = 6.300 S ; Rg =  .070 Ohm; Rs =    .070  Ohm
Cg = 447.0 pF; Cgd= 18.50 pF ; Cd =   168.0  pF
Lg = 1.600 nH; Ls =  .720 nH ; Ld =   1.300  nH
Rd =.270  Ohm; Cs = 3.00  pF
Pl =157.00 W ; Vd = 46.60 V  ; Rgs= 1.00E+12 Ohm
Lgd=   0.0 nH; Cn =   0.0 pF ; Ph =     0     -
Dg = -.40  dB; Rgd= 1.00E+12 Ohm
```

BLF177 Vds= 50 V Po= 150 W Class-AB

f	*	G	*	Inp.Imp.	*	Load Imp.
MHz	*	dB	*	Ohm	*	Ohm
1.6	*	54.70	*	2.29 - j133.58	*	6.23 + j .07
2.5	*	50.82	*	2.29 - j 85.50	*	6.23 + j .12
3.5	*	47.90	*	2.29 - j 61.09	*	6.23 + j .16
5.0	*	44.80	*	2.29 - j 42.78	*	6.22 + j .23
7.0	*	41.88	*	2.29 - j 30.58	*	6.21 + j .32
10.0	*	38.78	*	2.29 - j 21.45	*	6.18 + j .46
14.0	*	35.85	*	2.29 - j 15.37	*	6.13 + j .64
20.0	*	32.75	*	2.29 - j 10.84	*	6.03 + j .89
24.0	*	31.17	*	2.29 - j 9.09	*	5.95 + j 1.05
28.0	*	29.83	*	2.29 - j 7.85	*	5.85 + j 1.20

Fig.9 Calculated gain impedances of the BLF177.

By adding a gate-source resistor of 6.25Ω the power gain reduces from 29.8 to 23.3 dB at 28MHz.

2.3.2. Input matching circuit

As mentioned in section 2.1. a special circuit matches the input impedance of each transistor to the 6.25Ω of the input transformer. The matching network chosen can be treated as the half of a double π-section as described in Ref.2. Removing the in-and output capacitance the circuit changes in a T-section with C_i as capacitor and 2 inductances with a value of half the inductances of the double π-section (see Fig.10).

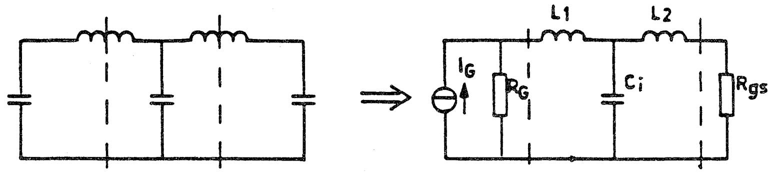


Fig.10 Input matching circuit.

C_i represents the input capacitance of the BLF177 and can be calculated from the input impedance of Fig.9.
For 7MHz: C_i=1/(2π*7E+6*30.58)=744pF.
Across this capacitor a constant voltage versus frequency from 1.6 up to 28MHz has to be developed. Provided C_i is an ideal capacitance the dimensioning of this network is as follows:
R_G=R_{gs} must be appr. 6Ω to obtain low I.M. distortion and good stability. This appeared during the development of the narrow band testcircuit as given in the BLF177 publication data. To judge whether this value is also acceptable for wideband operation we calculate the product:
$\omega_c.C_i.R_{gs}$ in which ω_c is the maximum angular frequency.
Doing so we find:
2π*28E+6*744E-12*6.25=0.818.
R_{gs} has been chosen 6.25Ω for the ease of transformation. Comparing the value of this product with the one given in Ref.2 we see that with a double π-section we can easely reach a bandwith of 50MHz. Therefore we have simplified the network as decsribed above.
Continuing the calculation we find:
L=0.997 R_G/ω_c=35.4nH (So L_1=L_2=17.7nH).
With the computer model mentioned in section 2.3.1. a gain of 22.3dB has been calculated with R_{gs}=6.25Ω.

Starting from this 22.3dB gain, $L_1=L_2=17.7nH$ and $R_{gs}=6.25\Omega$ the input VSWR and gain deviation have been calculated (see Fig.11).

INITIAL RESULTS
Rs = 6.250 Ohm; Gs = 22.300 dB
Par.LR : L= 17.700 nH; R = 6.250 Ohm
Ser.Ind. : L= 17.700 nH

f(MHz)	VSWR	dG(dB)
1.6	1.010	1.440
2.5	1.016	1.435
3.5	1.023	1.431
5.0	1.032	1.419
7.0	1.046	1.402
10.0	1.068	1.354
14.0	1.100	1.265
20.0	1.161	1.075
24.0	1.214	.910
28.0	1.281	.703

Fig.11 Results before optimization.

Before optimization the maximum VSWR=1.28 and the gain =22.7dB ± 0.37dB.
To achieve a maximally flat gain and a low input VSWR a computer optimization program has been used. This optimization results in a gain of 23.3dB with a maximum ΔGain = ± 0.09dB and a VSWR ≤1.09 (see Fig.12). For these results L_1 has been changed from 17.7nH to 9nH and L_2 from 17.7nH to 21.1nH. The R_{gs} has been decreased from 6.25Ω to 5.7Ω.

FINAL RESULTS
Rs = 6.250 Ohm; Gs = 23.300 dB
Par.LR : L= 21.079 nH; R = 5.749 Ohm
Ser.Ind. : L= 8.950 nH

f(MHz)	VSWR	dG(dB)
1.6	1.087	.072
2.5	1.087	.071
3.5	1.087	.074
5.0	1.086	.076
7.0	1.085	.084
10.0	1.082	.087
14.0	1.076	.086
20.0	1.059	.056
24.0	1.047	.006
28.0	1.044	-.087

Fig.12 Results after optimization

2.3.3. Input transformer

The input transformer is simular to the output transformer. It transforms the asymmetrical system impedance to the 2*6.25Ω=12.5Ω symmetrical source impedance. However the lower power handling (<3W) justifies a toroid of 4C6 material with smaller dimensions: 14x9x5mm (Dxdxh) which gives a volume A.l=0.445E-6m^3. As described in section 2.2.2. the primary winding can be calculated for L=20μH.

n_{pr}=SQR((L.l)/(μ_o.μ_r.A))

n_{pr}=SQR((20E-6*3.55E-2)/(4πE-7*120*12.54E-6

=19.4 turns. With n_{pr}=20 turns and a transformation ratio of 4:1 the n_{sec}=10 turns. V_{max}=SQR(2.P_i.R_S)=17.3V and B_{max}=0.013T.

The needed core volume A.l is:

A.l=(V_{max}/(ω.B_{max}))^2.(μ_o.μ_r)/L

A.l=(17.3/(2π*1.6E+6*0.013)^2.(4πE-7*120)/

20E-6=0.14E-6m^3.

The core used has a volume of 0.445E-6m^3. Fig.13 on page 16 shows the input transformer. The secondary winding has 10 turns of copper-foil (width 2mm, thickness 0.05mm). The primary winding has 20 turns of enamelled copper wire (0.5mm). Each secondary turn has been covered with 2 primary turns with a PTFE foil of 0.1mm thickness as isolation between the 2 windings. The method of winding is the same as described for the output transformer in section 2.2.2. The measured inductance is 20.95μH and L_{str}=250nH.
The correction method used for the input transformer is the same as described already in section 2.2.2. (see Fig.14).

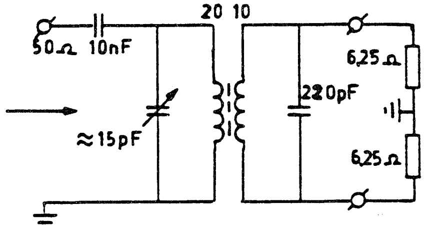


Fig.14 Inputtransformer with correction.

The transformer has been corrected with parallel capacitors for the higher frequencies and a series capacitor for the lower frequencies. Fig.15 on page 17 gives the return losses before and after the correction.

2.3.4. Tuning of the inputcircuit

For the practical tuning of the inputcircuit each transistor has been adjusted at V_{DS}=50V and a quiescent current of 0.5A.
The gain and input return losses have been measured in the frequency range 1.6 up to 35MHz. The best results have been achieved by changing the secondary correction capacitor of the inputtransformer from 220 to 30pF and the primary correction trimmer from ≈ 15pF to ≈20pF. The low frequency correction capacitor at the input has been removed. The inductance in serie with R_{gs} has been increased from 21.6 to 35nH. Fig.16 on page 18 gives the complete circuit diagram of the wide band amplifier with 2 BLF177 transistors. Fig.17 on page 19 gives the corresponding parts list.

3. CONSTRUCTION OF THE AMPLIFIER

For the printed circuit board double Cu-clad epoxy fibre glass has been used with a thickness of 1/16" and ε_r=4.5. The position of the components is on one side and the other side serves as a groundplane. Connections to the groundplane have been made with rivets and with straps under the source leads and at the edges of the PC-board on the in-and output side.
The printed circuit board has been attached to a solid copper plate (145x120x10mm) which functions as a heatsink. Around the position of both transistors a tube has been soldered in the copper plate to controle the temperature by means of a watercooling system. For a good thermal contact between heatsink and transistors heatsink compound has been used.
Fig.18 on page 20 shows the lay-out of the amplifier. The transformers have been fastened above the printed circuit board by means of accessories of Delrin material.
These accessories have been attached through the PC-board in the copper plate.

4. MEASURED PERFORMANCE

4.1. Single tone measurements

Fig.19 to 23 on page 21 and 22 show at a constant outputpower of 300W at 2 heatsink temperatures the gain, efficiency, input return losses, 2nd and 3rd harmonics at the output as a function of the frequency. In the range 1.6-28MHz the gain is 22-23dB, the efficiency 52.5-61%, the input return losses are below -15.5dB, the second harmonics better then -25dB and the third harmonics below -16dB.
At a heatsink temperature of 70°C the gain decreases about 1.5dB. The heatsink temperature has only little influence on the other parameters. Fig.24 to 26 on page 23 shows at 4 frequencies the output power as a function of the input power and the gain and efficiency versus outputpower.
Above 10MHz the efficiency decreases about 6%. At 20MHz the gain decreases above P_o=200W. At other frequencies this decrease starts at P_o=300W.

4.2. Two tone measurements

The two tone measurements have been carried out with 2 carriers with a frequency distance of 1KHz. Fig.27 to 30 on page 24 and 25 give as a function of the frequency the gain, efficiency, 3rd order distortion and 5th order distortion at 4 output levels. Over the whole frequency range the gain variation is less than 1dB at each power level. At P_o=300W PEP the efficiency is at least 40%, the 3rd order distortion $\leqslant$-33dB and the 5th order distortion $\leqslant$-38dB.
Fig.31 and 32 on page 26 give the 3rd and 5th order distortion versus output power at 4 frequencies.
To verify the choise of I_{DQ}=1A the 2nd and 3rd order distortion have been measured versus I_{DQ}. These measurements have been carried out at the most critical frequency and output level of 20MHz and 30W PEP resp.
Fig.33 on page 27 shows that I_{DQ}=1A for both transistors together was a good choise.

5. BALANCED CIRCUIT

As shown in the table below there is a certain amount of unbalance between both drain currents at RF operation. It is possible to improve this by using baluns in front of the input transformer and after the output transformer.

f (MHz)	I_{D1}(A)	I_{D2} (A)
1.6	5.2	4.8
5	5.1	4.75
10	5.25	4.8
15	5.3	4.95
20	5.7	5.3
25	6.2	5.25
30	5.85	5.15

Fig.34 Drain currents at P_o=300W.

6. CONCLUSIONS

This report shows that it is possible to design a wideband push-pull ampifier with 2 BLF177 MOS transistors having a very good performance.
The main properties are:

- Bandwidth 1.6-28MHz
- V_{DS} 50V
- I_{DQ} 1A
- Gain at P_o=300W 22-23dB
- Efficiency at P_o=300W 52.5-61%
- Return losses input at P_o=300W ≤-15.5dB
- 2nd harmonics output at P_o=300W ≤-25dB
- 3rd harmonics output at P_o=300W ≤-16dB
- IMD at P_o=300W PEP ≤-33dB

7. REFERENCES

1. M.J.Köppen
   Application report ECO 7308
   A single stage wideband (1.6-28MHz)linear power amplifier for 300W PEP using 2xBLX15.

2. G.Lukkassen
   Application report RNR-1-498-1986-AS
   A wideband power amplifier (25-110MHz) with the MOS transistor BLF245.

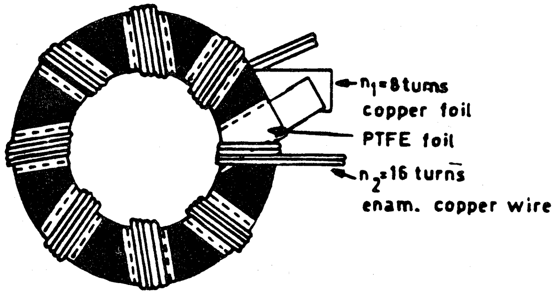


Fig.2

Output transformer

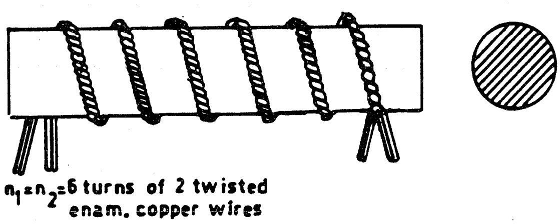


Fig.6

Tapped drain choke

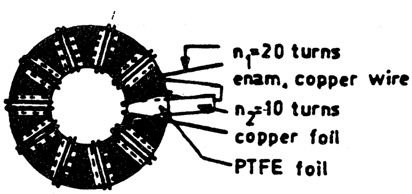


Fig.13

Input transformer

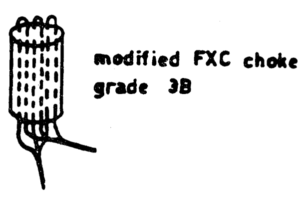


Fig.35

Decoupling choke

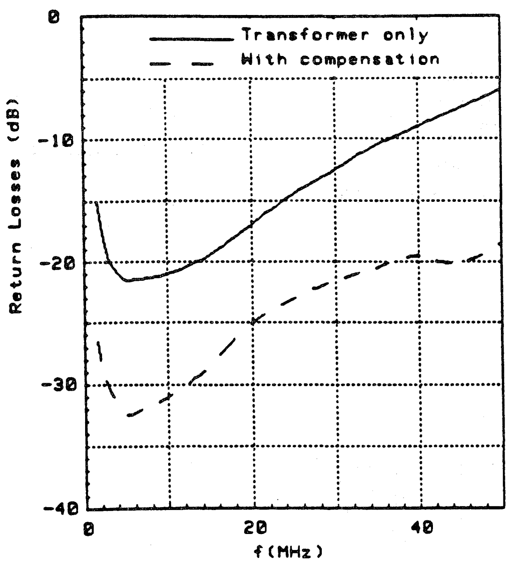


Fig.4 Output transformer correction

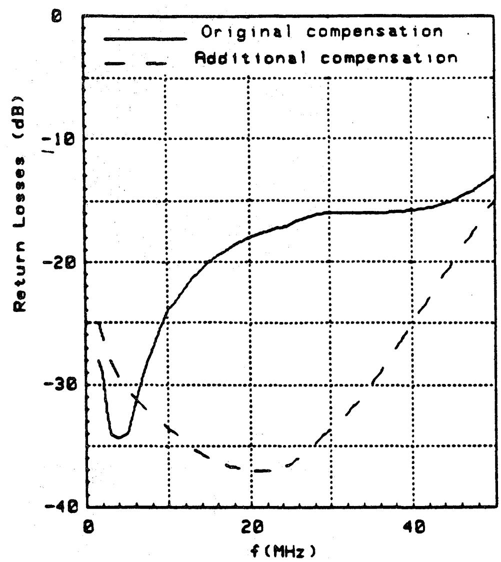


Fig.8 Return losses output circuit

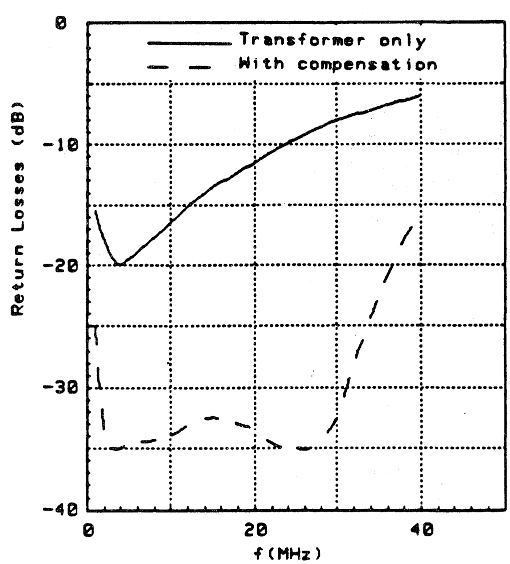


Fig.15 Input transformer correction

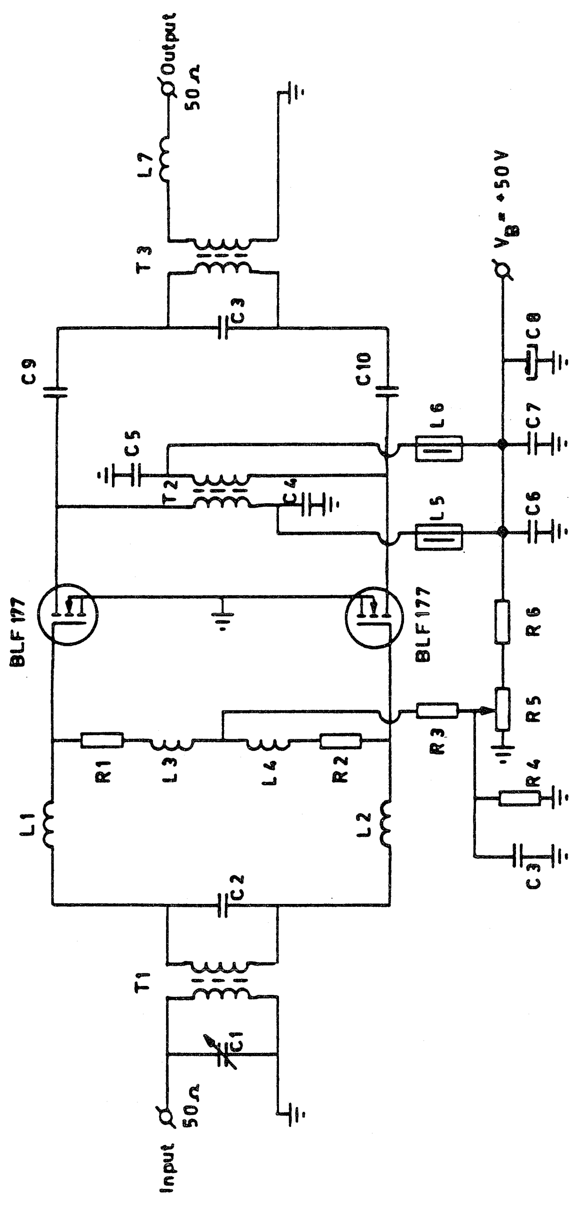


Fig.16 Circuit diagram of the 2x BLF177 amplifier

FIG.17 PARTS LIST OF THE WIDE BAND PUSH-PULL AMPLIFIER WITH 2xBLF177 (1.6-28MHz)

C1 = 5-60pF film dielectric trimmer (cat.nr.:2222 809 08003)
C2 = 30pF multilayer chip capacitor *
C3 = 2x100nF multilayer chip capacitor (cat.nr.:2222 852 47104)
C4 = C5 = 3x100nF metallized film capacitor (cat.nr.: 2222 368 21104)
C6 = C7 = 100nF multilayer chip capacitor (cat.nr.: 2222 852 47104)
C8 = 10μF (63V) electrolytic capacitor (cat.nr.2222 030 28109)
C9 = C10 = 4x10nF metallized film capacitor (cat.nr.: 2222 368 51103)
C11 = 2x75pF multilayer chip capacitor *
L1 = L2 ≈ 9nH, printed inductance; l=47mm,W=6mm
L3 = L4 = 35nH, 3 turns enamelled Cu-wire (0.7mm) int.dia.:3mm, l=2.35mm
L5 = L6 = 2.2μH, 1 turn through modified Ferroxcube choke grade 3B (cat.nr.:4312 020 36642) See Fig.35 page 16
L7 = 100nH, 5 turns enamelled Cu-wire (0.8mm) int.dia.: 5mm, l=6.1mm
R1 = R2=5.9Ω; 4 metal film resistors of 23.7Ω (0.4W) in parallel (cat.nr.: 2322 151 72379)
R3 = 1KΩ, metal film resistor (0.4W) (cat.nr.:2322 151 71002).
R4 = 1MΩ, metal film resistor (0.4W) (cat.nr.: 2322 151 71005).
R5 = 500Ω, Cermet potentiometer (0.75W)
R6 = 5.6KΩ, metal film resistor (1W) (cat.nr.: 2322 153 55622)
T1 = Input transformer:
n_{pr}=20 turns enamelled Cu-wire (0.5mm)
n_{sec}=10 turns copper foil (width 2mm, thickness 0.05mm) wound around toroidal core, grade 4C6, dimensions: 14x9x5mm (cat.nr. 4322 020 97181) see Fig.13 page 16
T2 = Drain choke:
6 turns of twisted pairs of 0.8mm Cu-wires (each winding consists of 2 wires in parallel) wound on a Ferroxcube rod, grade 4B1, dimensions 10x50mm, see Fig.6 on page 16.
T3 = n_{pr}=8 turns copper foil (width 6mm,thickness 0.05mm)
n_{sec}=16 turns of 2 enamelled Cu-wires (0.6mm) in parallel wound around toroidal core, grade 4C6, dimensions:36x23x15mm (cat.nr.4322 020 97201) see Fig.2 page 16.
2 of these transformers in parallel from the complete outputtransformer.

PC-board: double Cu-clad, 1/16" epoxy fibre glass (ε_r=4.5)
* American Technical Ceramics type 100B or capacitor of same quality.

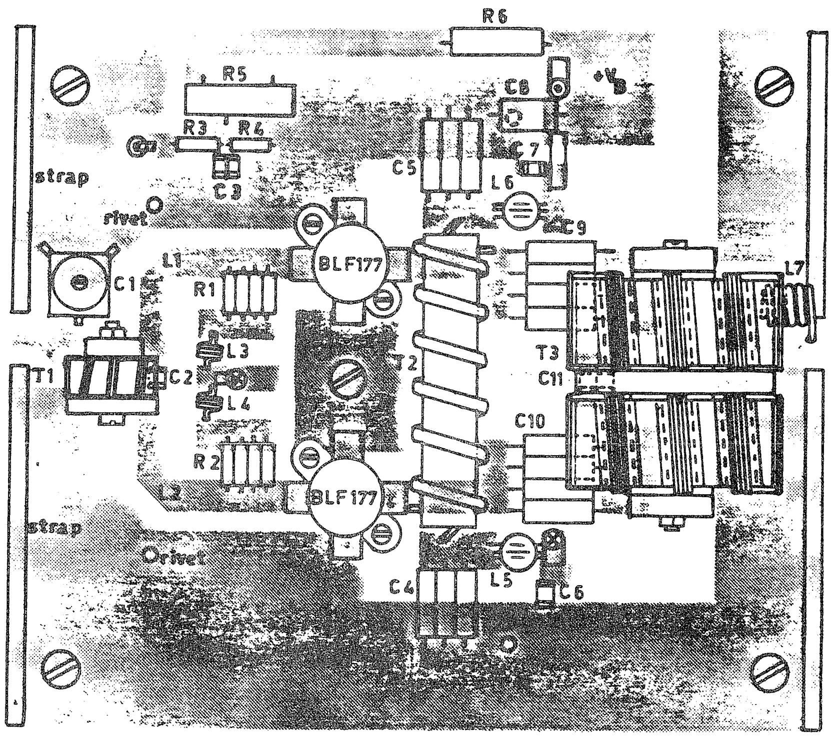


Fig.18 Lay-out of the 2x BLF177 amplifier

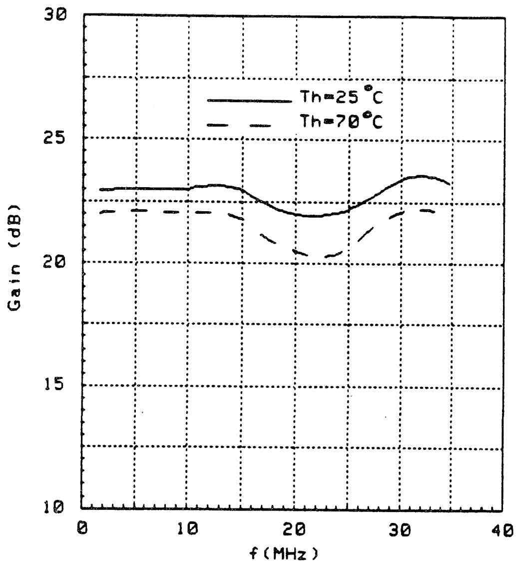


Fig.19 Gain versus frequency

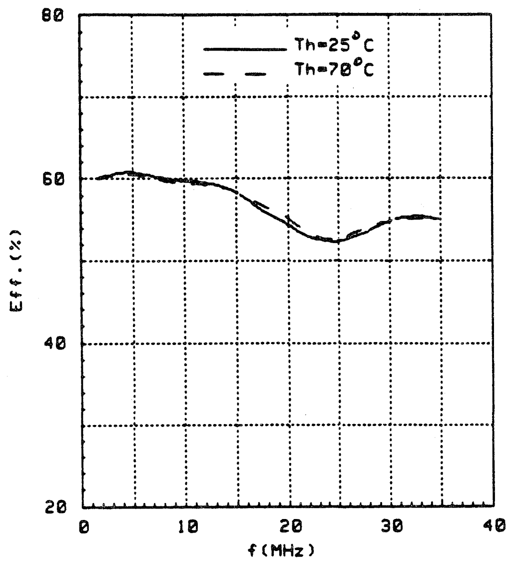
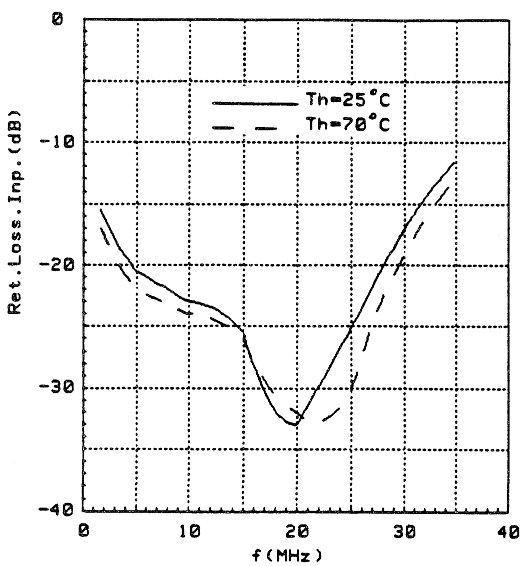


Fig.20 Efficiency versus frequency



2x BLF177

V_{DS}=50V

I_{DQ}= 1A

P_o =300W

Fig.21 Input return losses versus frequency

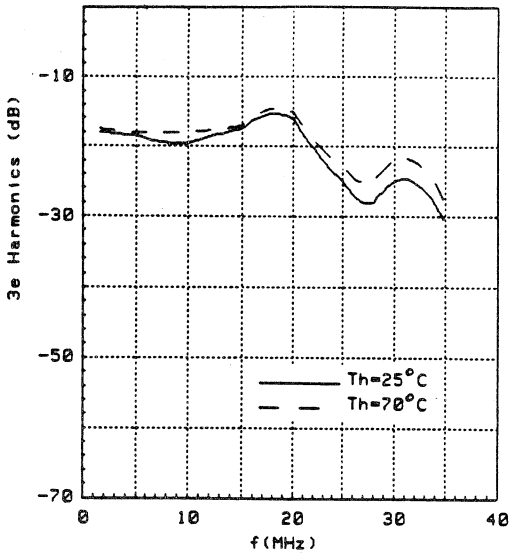


Fig.22 3rd harmonics versus frequency

2xBLF177

V_{DS}=50V

I_{DQ}=1A

P_o=300W

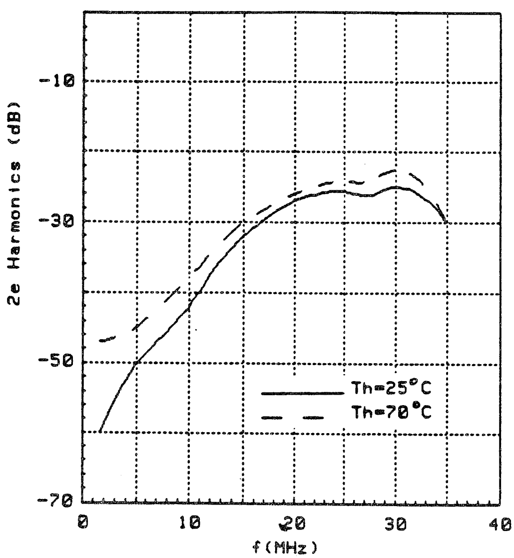


Fig.23 2nd harmonics versus frequency

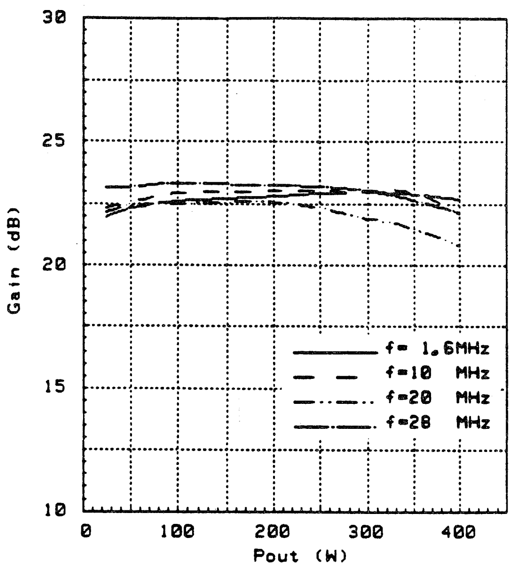


Fig.24 Gain versus outputpower

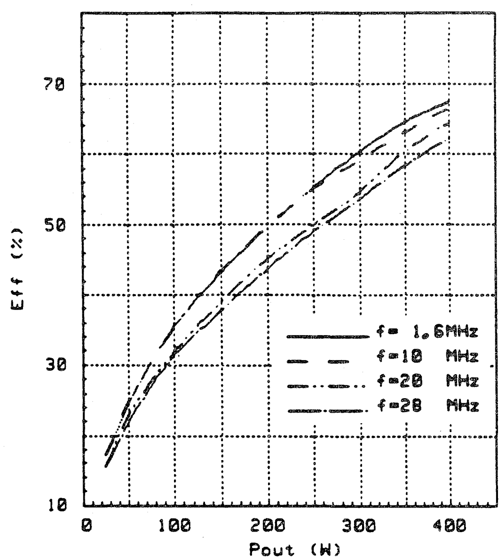


Fig.25 Efficiency versus outputpower

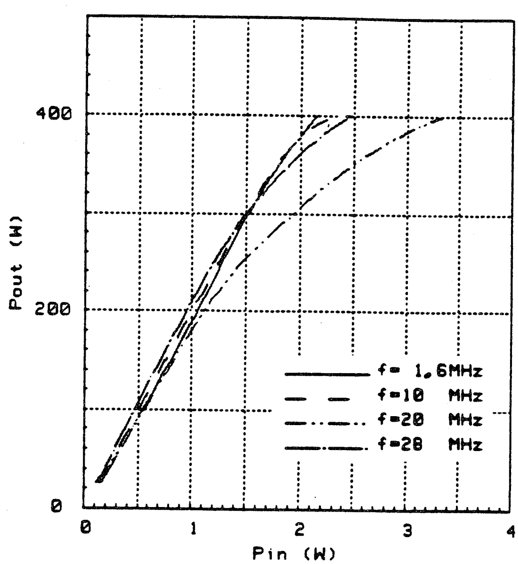
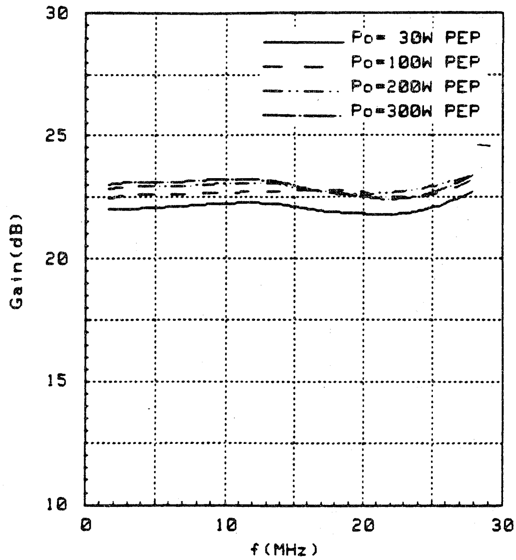


Fig.26 Outputpower versus inputpower

2x BLF177

V_{DS}=50V

I_{DQ}=1A



2x BLF177

V_{DS}=50V

I_{DQ}= 1A

$f_p - f_q$=1 KHz

Fig.27 Gain versus frequency

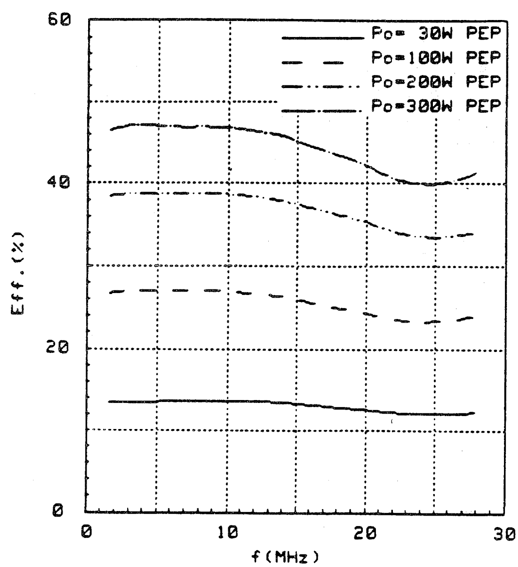


Fig.28 Efficiency versus frequency

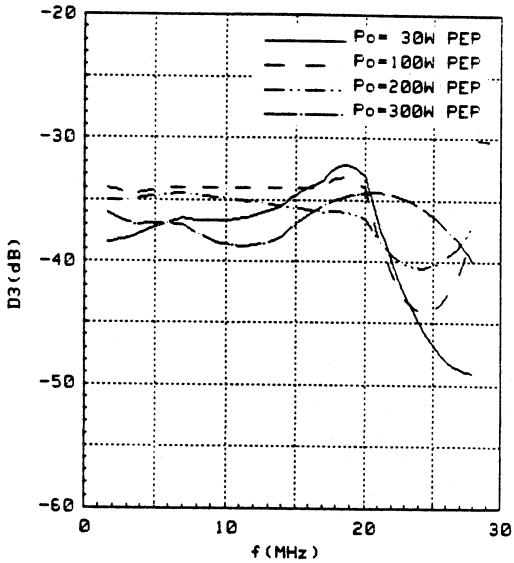


Fig.29 Third order dist. versus frequency

2x BLF 177

V_{DS}=50V

I_{DQ}= 1A

$f_p - f_q$= 1 KHz

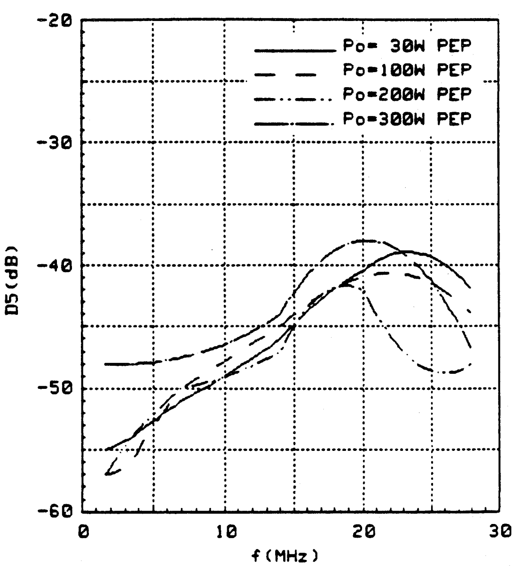
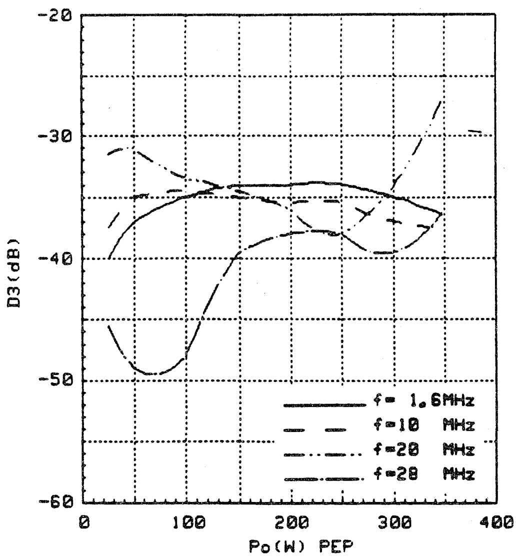


Fig.30 Fifth order dist. versus frequency



2x BLF177

V_{DS}=50V

I_{DQ}= 1A

$f_p - f_q$=1 KHz

Fig.31 3rd order distortion versus P_o

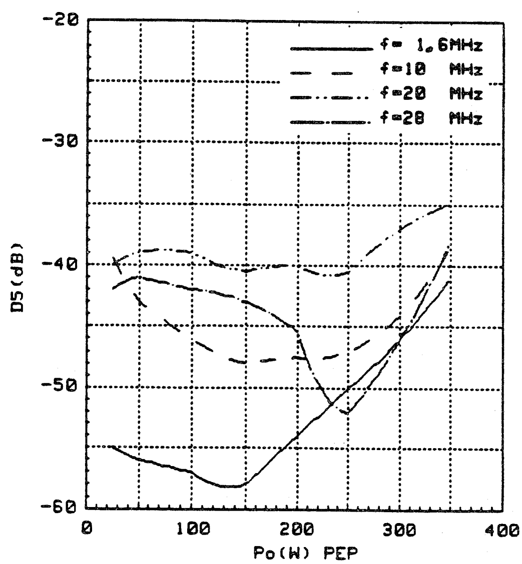


Fig.32 5th order distortion versus P_o

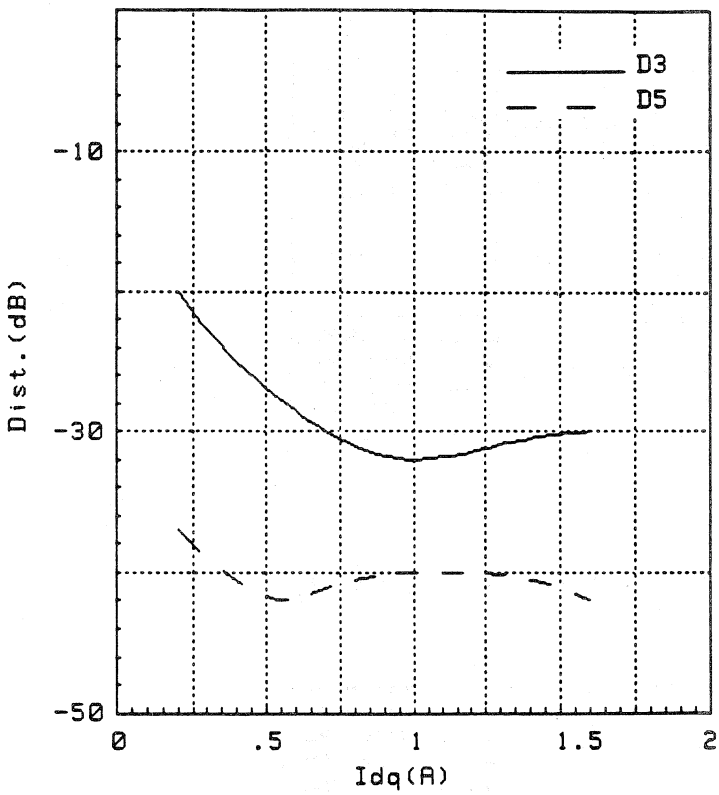


Fig.33 Distortion versus quiescent current

2x BLF177

V_{DS}=50V

f_p-f_q = 1 KHz

f_p= 20MHz

P_o= 30W PEP

Report no: RNR-1-358-1987 / NCO 8704
Author : R. Gajadharsing
Date : 1987-06-18

LINEAR PERFORMANCE OF BLF244 IN S.S.B. CLASS-A OPERATION

1. INTRODUCTION

This report contains results of measurements carried out on the BLF244 in S.S.B. class-A operation.
Linear measurements have been performed on six transistors from batch RI50C (ass.nr. 3030). Each transistor was taken from a different slice.

2. TESTCIRCUIT

Measurements have been done in a wideband amplifier designed for the frequency range 1.6-28MHz. The circuit diagram and component list are given on page 5.
Negative feedback (R2) has been employed to attain a flat gain of the amplifier. A shunt resistor(R1) between gate and source takes care of stable operation and also decreases the input resistance to 12.5 Ohm. Matching to 50 Ohm is accomplished with a 4:1 broadband transformer.

At the output side a broadband load of 50 Ohm is provided to the transistor.
A more detailed decription of this kind of amplifiers is given in application report RNR-1-376-1987.

3. TESTCONDITIONS

The quiescent drain current for class-A operation is set to 0.6A at a supply voltage of 28V. This is below the maximum allowable DC-current for a heatsink temperature of 70 °C which is 0.9A for this device.

Linearity measurements have been performed with two tones of equal amplitude with a frequency seperation of 1kHz.
The intermodulation distortion products d3 and d5 are referred to the amplitude of one of the two tones.

The transistors have been tested at a nominal output power of 4W PEP, with a heatsink temperature of 25 °C.

4. TESTRESULTS

The table below contains results of measurements at f=28MHz of 6 devices.

Conditions: Vds=28V; Idq=0.6A; Pout=4W PEP; Ths=25 °C.

BATCH RI50C (ass.nr.3030)

Dev.no.- Sliceno.	Pin (mW)	Gp (dB)	d3 (dB)	d5 (dB)	input ret.loss (dB)
2- 2	9.0	23.5	-40.5	-60	-20.5
18- 3	9.0	23.5	-41.0	-60	-22.0
27- 10	8.8	23.6	-40.5	-60	-24.5
32- 12	8.9	23.5	-40.5	-60	-22.5
44- 19	8.8	23.6	-40.5	-60	-24.0
52- 21	8.8	23.6	-40.5	-60	-23.0

Measurements have also been performed versus output power at f=28MHz. Fig.1 and 2 on page 3 show the powergain and IMD(d3) of a typical device (dev.no.52 from slice 21). Pout is varied between 0.5W and 8W P.E.P. wich resulted in a gain variation of approximately 0.5dB. IMD(d3) exceeds the level of -40dB for an output power greater than 4.3W P.E.P..
The amplifier performance versus frequency has also been measured at Pout=4W P.E.P. with the same device. Fig.3, 4 and 5 on page 4 show the powergain, IMD(d3) and input return loss versus frequency. The measuring frequency extends from 1.6MHz to 32MHz. The resulted powergain is 24dB +/-0.4dB and IMD(d3) varies between -48dB and -40.5dB while the input return loss is better than -20dB.

5. CONCLUSION

The BLF244 is suited for linear operation in Class-A in the HF-band. It has an IMD(d3) of better then -40dB up to an output power of 4W P.E.P. throughout the band at Vds=28V and Idq=0.6A.

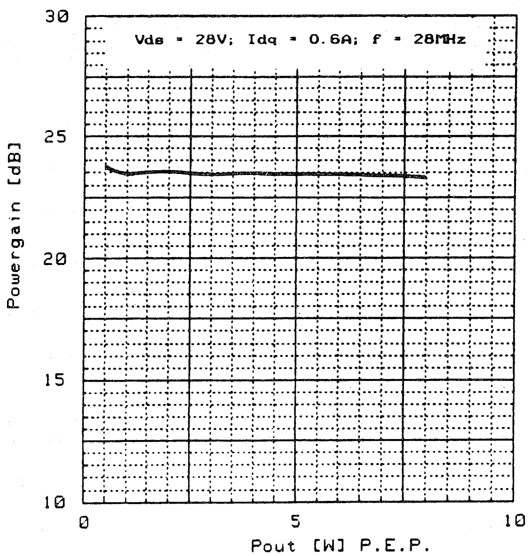


Fig.1 Powergain versus Pout

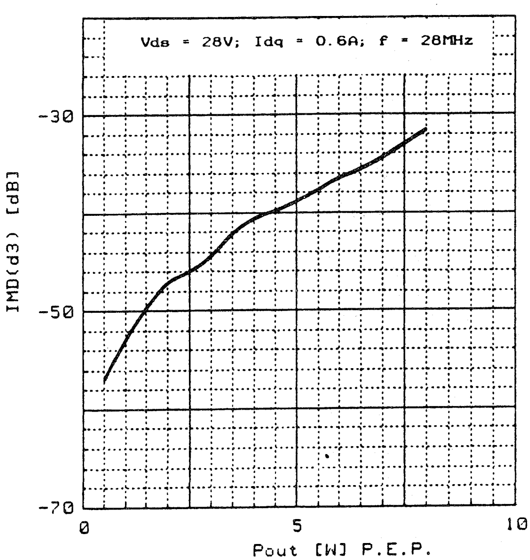


Fig.2 IMD(d3) versus Pout

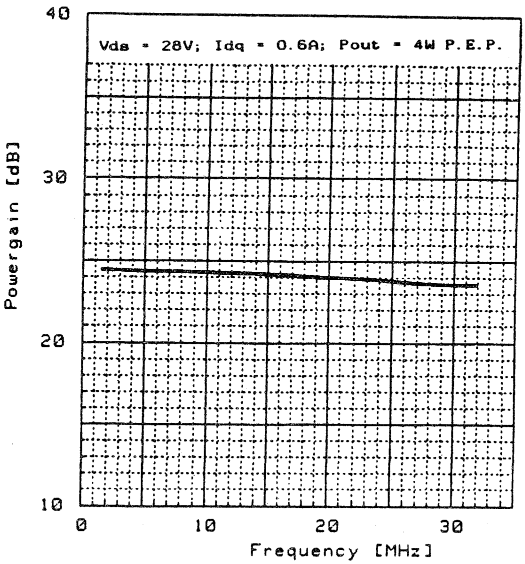


Fig.3 Powergain versus frequency

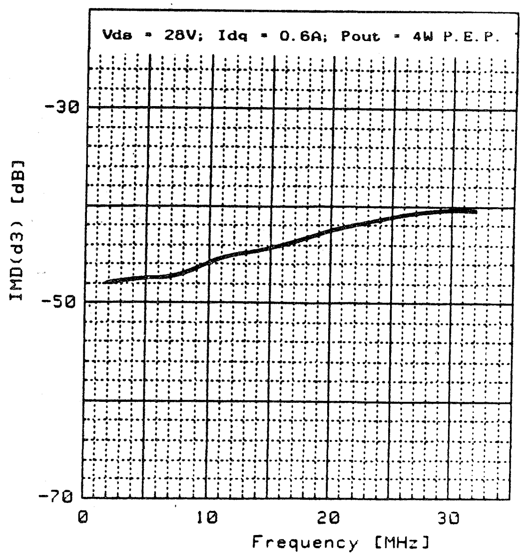


Fig.4 IMD(d3) versus frequency

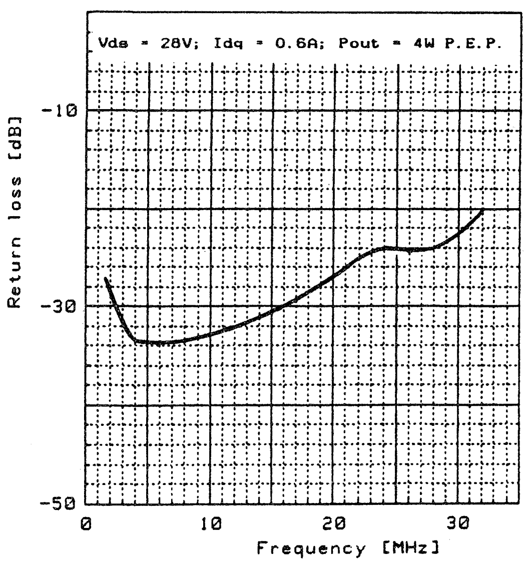
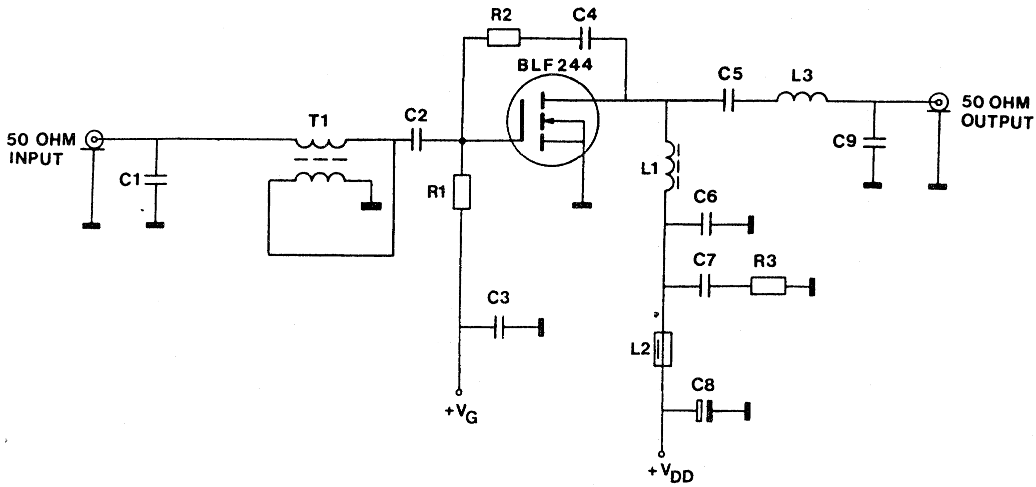


Fig.5 Input return loss versus frequency



Circuit diagram of the wide band amplifier for BLF244

LIST OF COMPONENTS

Capacitors

C1= 3.9pF; multilayer ceramic chip capacitor *
C2= 3*10nF; multilayer ceramic chip capacitor
(cat.nr.2222 852 47103)
C3=C4=C6= 100nF; multilayer ceramic chip capacitor
(cat.nr.2222 852 47104)
C5= 10nF; multilayer ceramic chip capacitor
(cat.nr.2222 852 47103)
C7= 3*100nF; multilayer ceramic chip capacitor
(cat.nr.2222 852 47104)
C8= 10uF(63V); Aluminium electrolytic capacitor
(cat.nr.2222 030 28109)
C9= 24pF; multilayer ceramic chip capacitor *

Inductors

L1= 20uH; drain choke, 36 turns enamelled Cu-wire (0.7mm)
wound on a Ferroxcube rod grade 4B1,
dimensions(5*30)mm
L2= Ferroxcube RF choke, grade 3B (cat.nr.4312 020 36640)
L3= 189nH; 8 turns enamelled Cu-wire (1.0mm);
int.dia.= 5.0mm, length= 9.5mm; leads 2*3.0mm

Resistors

R1= 16 Ohm; metal film resistor; 0.4W
R2= 1500 Ohm; metal film resistor; 0.4W
R3= 10 Ohm; metal film resistor; 0.4W

Transformer

T1- 4:1 transformer; 18 turns of twisted pair of 0.25mm
enamelled Cu-wire (10 twists per cm) wound on a
toroidal core grade 4C6, dimensions (9*6*3)mm.
(cat.nr. 4322-020-97171)

Printed circuit board: double sided Cu-clad epoxy fibre
-glass laminate (ε_r=4.5),
thickness 1/16 inch.

* American technical ceramics capacitor type 100B.

Report no : RNR-1-376-1987-AS / NCO 8705
Author : R.Gajadharsing
Date : Augustus 1987

TITLE

A WIDEBAND LINEAR AMPLIFIER (1.6-28MHz) FOR 8W PEP IN CLASS-A WITH THE MOS-TRANSISTOR BLF175(V_{ds}=50 Volt)

SUMMARY

In this report a description is given of a wideband linear amplifier intended for driver applications in SSB transmitters for the frequency range 1.6-28MHz.
It employs a MOS-transistor BLF175 suited for a supply voltage of 50V.
The transistor is adjusted in class-A with a quiescent drain current of 800mA.
The main properties at P_o=8W PEP are:

Powergain	:	28.3 - 28.6dB
IMD (d3)	:	≤ -41dB
IMD (d5)	:	≤ -60dB
Input return loss	:	≤ -26dB

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1. **INTRODUCTION**

The amplifier that will be discussed in this report concerns a wideband linear amplifier, designed for driver applications in SSB transmitters in the HF band.
This design is based on the RF power MOS-transistor BLF175 which is primarily designed for communication purposes in the HF-band. This device can deliver 8W PEP in class-A at an IMD (d3)<-40dB, when operated from a supply voltage of 50 Volt.
It is encapsulated in a SOT123 four-lead flange type with a ceramic cap.

2. **GENERAL CONSIDERATIONS**

One of the most important factors to be considered in the design of driver stages for SSB transmitters is intermodulation distortion. The major cause for intermodulation distortion is the non-linear transfer characteristic of the transistor. A generally accepted IM distortion figure is <-40dB. To achive this, driver stages must be operated in class-A. One of the properties of a class-A amplifier is its low efficiency, which for pre-drivers is of less importance.

The amplifier must have a flat gain response, within a few tenths of a dB. Its response should preferably be superiour to that of the final amplifier of a SSB transmitter.
The input return loss versus frequency must be low because, it will possibly form the load of a pre-driver.

3. **DESIGN OF THE AMPLIFIER**

3.1. Circuit description
Fig.1 shows the basic circuit of this broadband amplifier. Negative feedback combined with parallel input compensation has been applied to obtain flat gain and low input return loss.

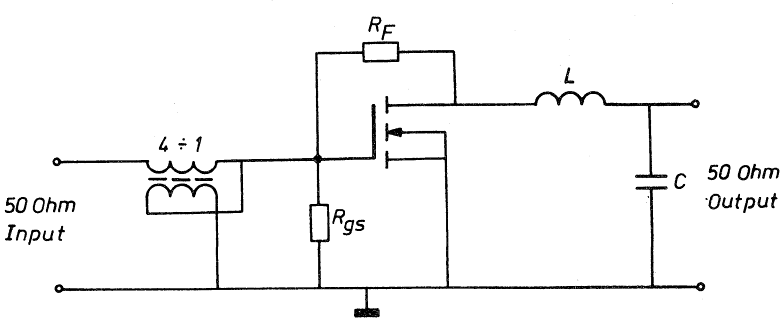


Fig.1 Basic circuit of the wideband amplifier.

Matching of the input to 50Ω is accomplished with a 4+1 broadband transformer of the transmission line type. At the output side a LC-section compensates the output capacitance of the transistor for the frequency range of interest in order to provide the transistor with a constant resistive load.

3.2. Design procedure
The amplifier will be designed for a supply voltage of 50 Volt and a system impedance of 50Ω.

First the DC-operating point must be determined. The most important factor that restricts the DC-current in MOS-transistors is the maximum allowable power dissipation in the transistor. For a maximum operating junction temperature of 200°C and a maximum allowable heatsink temperature of 70°C the maximum dissipation with R_{thj-h}=2.9 K/W equals to 44.8W. This corresponds with a drain current of 0.9A at V_{ds}=50 Volt. In order to keep the dissipation within safe limits I_{ds} is set to 0.8A.

Second the optimum load resistance is determined. For class-A amplifiers this is given by the relation:

$$R_L = \frac{V_{ds}}{I_{ds}} \qquad (1)$$

In this case R_L equals to 50/0.8=62.5Ω. In order to avoid an output transformer R_L is chosen to be 50Ω.

Now the load resistance has been established, the input resistance can be determined. This resistance is formed by the input shunt resistance and that part of the feedback resistance reflected to the input.
Several properties of this amplifier are determined by this resistance, viz.:
a. the power gain
b. the cut-off frequency.
In the next sections a brief analysis will be given of this amplifier in order to determine the input resistance and the powergain.

3.2.1. Powergain

For class-A amplifiers small signal analysis produces sufficiently accurate results. The small-signal equivalent circuit of the amplifier is shown in fig.2. All transistor package parasitics are neglected for this frequency range.

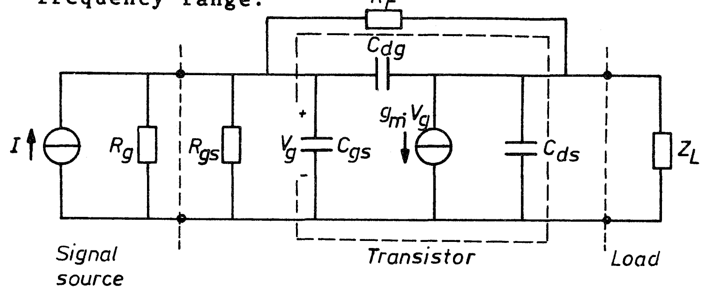


Fig.2 Small signal equivalent circuit of amplifier.

The Y-parameters of the transistor with feedback resistor R_F and shunt resistor R_{gs} are:

$$Y_{11} = G_{gs} + G_F + j\omega(C_{gs} + C_{gd}) \tag{2}$$

$$Y_{12} = -j\omega C_{gd} - G_F \tag{3}$$

$$Y_{21} = g_m - G_F - j\omega C_{gd} \tag{4}$$

$$Y_{22} = G_F + j\omega(C_{ds} + C_{dg}) \tag{5}$$

The general expression for powergain of any linear amplifier is:

$$G_p = \frac{P_o}{P_i} = \frac{G_L\,|Y_{21}|^2}{|Y_{22} + Y_L|^2 \cdot Re(Y_{in})} \tag{6}$$

in which:

$$Y_{in} = Y_{11} - \frac{Y_{12} \cdot Y_{21}}{Y_{22} + Y_L} \tag{7}$$

The load admittance is:

$$Y_L = G_L - j\omega(C_{ds} + C_{dg}) \tag{8}$$

After substitution of equation (2)-(5) and (7), (8) into (6) we obtain:

$$G_p = \frac{G_L[(g_m - G_F)^2 + \omega^2 C_{gd}^2]}{(G_L + 2G_F)[G_{gs}(G_L + 2G_F) + G_L G + G_F^2 + \omega^2 C_{gd}^2]} \tag{9}$$

in which:

$$G = G_F(1 + \frac{g_m}{G_L}) \tag{10}$$

If G_F and ωC_{gd} are assumed very small with respect to G_L and G_m we get the simple expression:

$$G_p = \frac{g_m^2}{G_L(G_{gs} + G)} \tag{11}$$

3.2.2. Cut-off frequency

The cut-off frequency of this amplifier is dominated by the input circuit. The output circuit has a much higher cut-off frequency and is therefore not relevant.

Fig.3 shows the unilaterised small-signal equivalent circuit of Fig.1.

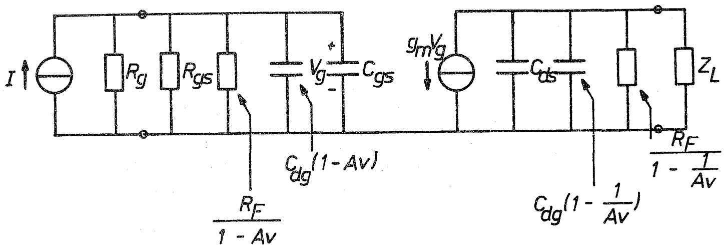


Fig. 3 unilaterised small-signal equivalent circuit.

Av is the voltage gain between drain and gate which is assumed to be a real number. The total input resistance is:

$$R_i = R_{gs} // \frac{R_F}{1 - Av} \tag{12}$$

And the input capacitance is:

$$C_i = C_{gs} + C_{gd}(1 - Av) \tag{13}$$

The 3dB cut-off frequency of this RC-combination is given by:

$$f_c = \frac{1}{2\pi R_i C_i} \tag{14}$$

So, if Ci is known we can determine Ri for a certain bandwidth.

3.3. <u>Calculations</u>

Calculations are based on transistors from one batch of the BLF175. The mean values of the transistor parameters were taken. These are:

g_m = 1.5 S (V_{ds}=10V; I_d=1A)

C_{gs} = 145.1 pF

C_{ds} = 34.4 pF (V_{ds}=50V; V_{gs}=0V; f=28MHz)

C_{gd} = 3.42pF

The transconductance (gm) of this device is determined by a pulse measurement. Under normal operating conditions gm will be lower due to the higher junction temperature. The reduction is approximately 25% for normal operating conditions. The effective transconductance is therefore:

$$g_{me} = 1.5 * 0.75 = 1.1 \text{ S}$$

The capacitors C_{ds} and C_{gs} are voltage-dependent. Due to RF-excitation the effective capacitance in class-A will be 10% higher, so:

$$Cds_e = 1.1 * C_{ds} = 37.8\text{pF}$$
$$Cgd_e = 1.1 * C_{gd} = 3.76\text{pF}$$

The voltage gain between drain and gate in Fig.1 can be calculated with:

$$A_v = -gm_e . R_L \qquad (15)$$

For a load resistance of 50Ω this amounts to:

$$A_v = -1.1 * 50 = -55$$

According to eq.(13) the total input capacitance amounts to:

$$C_i = 145.1 + 3.76 * (1+55) = 355.7\text{pF}$$

The total input resistance can now be determined with eq.(14). For f= 28MHz this amounts to:

$$R_i \leq \frac{1}{2 \times \pi \times 28 \times 10^6 \times 355.7 \times 10^{-12}} = 16 \text{ Ohm}$$

For the ease of transformation a value of 12.5Ω has been chosen. The cut-off frequency therefore increased to 35.8MHz.

R_i consists of the parallel connection of R_{gs} and $R_F/(1-A_v)$, see fig.3. First the feedback resistance R_F will be calculated.
The only restriction that holds for the feedback resistance is the power dissipation in it. This must be kept low in order to prevent deterioration in IMD performance.
The dissipation allowed, is set to ≈3% of the RMS output power. According to fig.3 the total reflected feedback resistance to the output side is:

$$R_F' = \frac{R_F}{1 - \frac{1}{Av}} \tag{16}$$

Because $A_v >> 1$, $R_F' \approx R_F$ and amounts to:

$$R_F = 1500\Omega$$

Now, R_{gs} can be calculated for R_i=12.5Ω.
With $R_F/(1-A_v)$=26.8Ω we find for R_{gs} a value of 23.4Ω. The closest practical value is 24Ω.

The powergain in dB can be determined with eq. (9), and is calculated to be:

$$G_p = 10 \log (636.6) = 28.4\text{dB}$$

When the approximate equation is used (11) we get:

$$G_p = 10 \log (756.3) = 28.8\text{dB}$$

So, a good estimation is obtained when eq. (11) is used.

3.4. Output matching

The output impedance of the transistor can be represented by a parallel connection of a resistance and a capacitance.
The resistance has a value of 50Ω, see section 3.2., and the capacitance is equal to
$1.1 * (C_{ds}+C_{dg})$=41.6pF due to R_F-excitation.
This output capacitance is compensated by a LC-section for the frequency range of interest in order to obtain a constant resistive load, see fig.3.

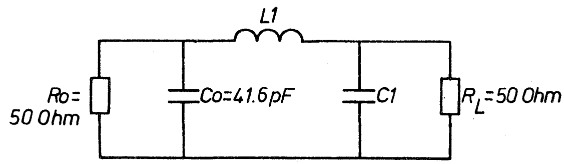


Fig.3 Output matching

According to Ref.[1] the component values of L_1 and C_1 for a cut-off frequency of 28MHz are:

L_1=189nH and C_1=41.6pF with VSWR=1.05

The output section contains two additional components, viz.:
1. a drain choke for biasing
2. a dc-blocking capacitor.

For RF-signals the drainchoke is connected in parallel with the output impedance, see fig.4, and must therefore be large enough, in order to avoid performance degradation at the low end of the band.
For the lowest frequency of interest (1.6MHz) the choke inductance must be at least:

$$L_{ch} = \frac{4 R_o}{2 \pi f_{min}} \tag{17}$$

In this case L_{ch} amounts to 20μH.
In practice this is obtained by winding 36 turns of enamelled copper-wire (0.7mm) on a ferroxcube rod, grade 4B1, with a length of 30mm and a diameter of 5mm. Because of the open magnetic circuit saturation due to DC-current will hardly occur.

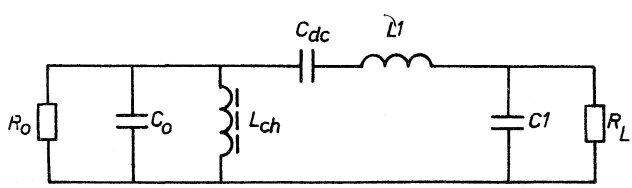


Fig.4 Output section.

The dc-blocking capacitor can be used to compensate the choke inductance for low frequencies.
According to [1] this capacitor must be 8nF for f=1.6MHz with $VSWR_{max}$=1.03.
In practice a chip capacitor was used of 10nF.

The matching performance of the output section was verified with an impedance analyzer. The transistor was first replaced by a dummy transistor which consisted of a resistor of 50Ω and a capacitor of 42pF in a SOT123 header. The return loss was measured at the load connection.
It appeared that the return loss improved when C_1 was replaced by a capacitor of 24pF. This was due to parasitics introduced by the printed circuit board and the additional components like the drainchoke and the blocking capacitor.
The return loss was better than -20dB throughout the band.

3.5. Input matching
The input section is shown in Fig.5.

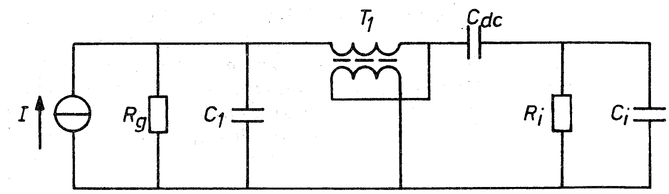


Fig.5 Input section.

A 4:1 broadband transformer is applied of the transmission line type. It utilizes a twisted-wire-pair transmission line wound on a toroidal core.
The windings are uniformly distributed around the toroid. Fig.6 shows the electrical circuit diagram and constructional details of this transformer.

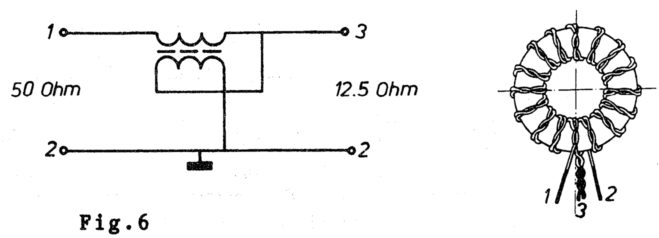


Fig.6

The required characteristic impedance of the transmission line is:

$$Z_o = \sqrt{R_g \cdot R_i}$$

For this case Z_o equals to $\sqrt{50*12.5} = 25\Omega$.
In practice Z_o will deviate from the required value and compensation will be necessary to improve the broadband performance of the transformer.
The characteristic impedance of 25Ω has been obtained by twisting two enamelled copper wires of 0.25mm-bare diameter. The wire diameter with isolation included was 0.27mm. Approximately 10 twists per cm were applied and the total wire length was 25cm.

A ferroxcube toroid, grade 4C6, has been applied with dimensions (9x6x3)mm. Here the size is not primarily determined by the power handling capabilities, but the required number of turns needed to establish the parallel inductance between the transformer terminals. On the other hand this inductance must not be higher than necessary, because the broadband performance of the transformer will degrade if the transmission line becomes longer than λ/8. A good practical value is that given by equation (17). This means for the inductance at the 50Ω side a value of 20μH and for the 12.5Ω side a value of 5μH. The number of turns needed is that which is required to make 5μH.
According to the design information in ref. [2], 13 turns for this toroid were required to make 5μH.
From measurements it appeared to be too low. Therefore the number of turns had to be increased to 18. This is due to deviation in material properties which for smaller toroids is larger.

The dc-blocking capacitor compensates the parallel inductance of 5μH. For 1.6MHz, 31.8nF is necessary according to ref. [1]. Three chip capacitors in parallel were used of 10nF each.

High frequency compensation for deviation in Z_o is accomplished by parallel capacitors between the transformer terminals. At the low ohmic side a part of C_i provides the required capacitance while at the high ohmic side C_1 provides this. Its value is determined by tuning a variable capacitor for optimum return loss at f=28MHz under nominal operating conditions. The required value was 3.9pF.

4. AMPLIFIER ALIGNMENT

The amplifier was constructed according to the design procedure given in the previous chapter.
Measurements were performed throughout the band at an outputpòwer of 8W PEP. The results are given below.

Powergain	= 27.5 - 28.6dB
IMD(d3)	$\leq$ -41dB
IMD(d5)	$\leq$ -60dB
Input return loss	$\leq$-18.5dB

The highest powergain occured at f=1.6MHz.
The total variation in gain of 1.1dB was found to be relatively large. In order to improve this compensation measures we considered.
There were two possibilities, viz.:
1. parallel input copensation; an inductance in series with the input shunt resistance which increases the effective shunt resistance at high frequencies and hence the gain.
2. feedback compensation; an inductance in series with the feedback resistance wich decreases the feedback at high frequencies and hence improves the gain.

The drawback of the latter is the relatively large inductance required for compensation, a few µH. The former is more elegant because of the low value of the required inductance.
Calculation of the optimum inductance for maximally flat response is complicated. Therefore its value was determined in an emperical way.
An inductance of 86nH was found, which reduced the total variation to 0.3dB for an average gain of 28.4dB.
An additional advantage of this compensation measure was the improvement of the input return loss, which became better than -26dB.

5. AMPLIFIER CONSTRUCTION

5.1. Construction notes

The circuit diagram and component list are given on page 17. The circuit board of this amplifier design is made of two-sided copper clad epoxy fibre-glass laminate with a thickness of 1/16 inch and a dielectric constant of 4.5. A full sized pattern of the printed circuit board is shown in fig.7 on page 18. The other side is fully metallized and used as ground plane. The ground planes on each side of the board are connected together by means of copper straps at the source leads and the N-connectors and the mounting screws.
Fig.8 on page 18 shows the component layout.
The unavoidable strip in the feedback path represents an inductance of 12nH and a capacitance of 5.5pF which can be neglected with respect to the feedback resistance. C4 is a dc-blocking capacitor and should have a low reactance for all frequencies.
To prevent low frequency spurious oscillation, a network comprising C7 and R3 is applied. At low frequencies R3 serves as a series loss for choke L2 and thus avoids a high Q factor.
C3 and C6 are small bypass capacitor for the carrier frequency.
L3 needs to be as large as possible and still be able to handle the required current. C8 must provide a solid bypass at all frequencies including the very low ones.

5.2. Heatsink

The circuit board is attached to a solid brass plate, which is provided with a circular hole for cooling purposes. A water cooling system controls the heatsink temperature.

6. AMPLIFIER PERFORMANCE

6.1. General

Performance measurements were carried out under the following conditions:

Supply voltage : V_{dd} = 50V
Quiescent drain current: I_{dq} = 0.8A
Heatsink temperature : T_{hs} = 25^{o}C

The measuring frequency extends from 1.6MHz to 32MHz. Two tones of equal amplitude were used with a frequency separation of 1KHz. The distortion products were measured with respect to one of the two tones.

6.2. Performance at constant output power

The measurements were done at an output power of 8W PEP.
The results obtained are:

Powergain = 28.3- 28.6dB, see fig. 9 on page 19;
IMD (d3) =-41.4- -47.4dB, see fig.10 on page 19;
IMD (d5) ≤-60dB;
Input return loss
≤- 26dB, see fig.11 on page 19.

6.3. Performance at constant frequency

As shown in fig.10 on page 19 the worst third order IMD product occurs at the highest end of the band. Therefore, measurements versus output power were only carried out at f=28MHz.
The results obtained are:

Powergain = 28.1- 28.4dB, see fig.12 on page 20;
IMD (d3) = -60 - -33.9dB, see fig.13 on page 20
-40dB is exceeded for $P_o \geq$ 9.5W PEP;
IMD (d5) ≤ -58dB;
Input return loss
≤ -21.5dB.

7. CONCLUSION

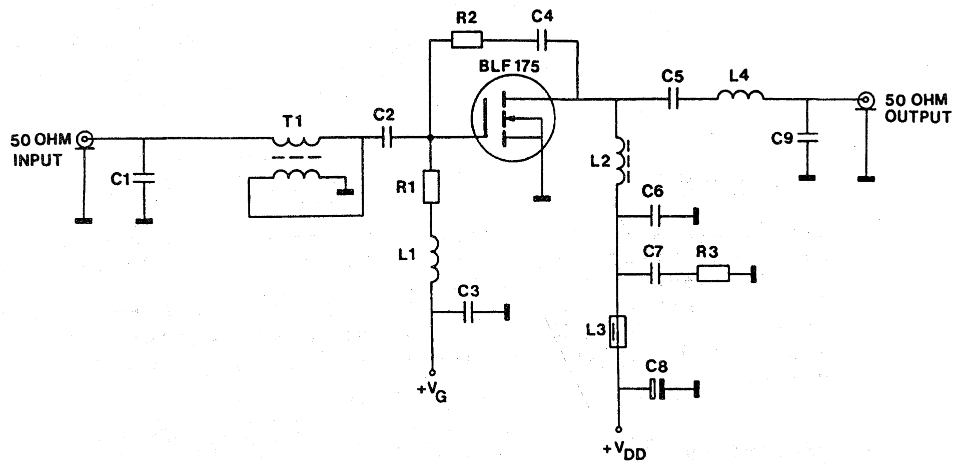
The design and construction of a wideband linear amplifier has been presented, with the MOS-transistor BLF175, for the frequency range 1.6-28MHz.
The transistor is adjusted in class-A and shows good linearity, IMD (d3) $\leq$ -40dB, up to an output power of 9.5W. It is suited for driver applications in SSB transmitters.

8. REFERENCES

[1] H.Nielinger; "Optimale dimensionering von Breitband-anpassungsnetwerken"; NTZ 1968, Heft 2, pp. 88-91.

[2] Philips Data Handbook; "Ferroxcube for power, Audio/Video and accelerators"; Book C5 1986, pp. 317.

Circuit diagram of the wideband linear amplifier.



LIST OF COMPONENTS

Capacitors

C1= 3.9pF; multilayer ceramic chip capacitor *
C2= 3*10nF; multilayer ceramic chip capacitor
(cat.nr.2222 852 47103)
C3=C4=C6= 100nF; multilayer ceramic chip capacitor
(cat.nr.2222 852 47104)
C5= 10nF; multilayer ceramic chip capacitor
(cat.nr.2222 852 47103)
C7= 3*100nF; multilayer ceramic chip capacitor
(cat.nr.2222 852 47104)
C8= 10uF(63V); Aluminium electrolytic capacitor
(cat.nr.2222 030 28109)
C9= 24pF; multilayer ceramic chip capacitor *

Inductors

L1= 86nH; 4 turns enamelled Cu-wire (0.6mm); int.dia.= 5.0mm,
length= 3.3mm; leads 2*2.0mm
L2= 20uH; drain choke, 36 turns enamelled Cu-wire (0.7mm)
wound on a Ferroxcube rod grade 4B1,
dimensions(5*30)mm
L3= Ferroxcube RF choke, grade 3B (cat.nr.4312 020 36640)
L4= 189nH; 8 turns enamelled Cu-wire (1.0mm); int.dia.= 5.0mm,
length= 9.5mm; leads 2*3.0mm

Resistors

R1= 24 Ohm; metal film resistor; 0.4W
R2= 1500 Ohm; metal film resistor; 0.4W
R3= 10 Ohm; metal film resistor; 0.4W

Transformer

T1- 4:1 transformer; 18 turns of twisted pair of 0.25mm
enamelled Cu-wire (10 twists per cm) wound on a
toroidal core grade 4C6, dimensions (9*6*3)mm.
(cat.nr. 4322-020-97171)

Printed circuit board: double sided Cu-clad epoxy fibre
-glass laminate (εr=4.5),
thickness 1/16 inch.

* American technical ceramics capacitor type 100B.

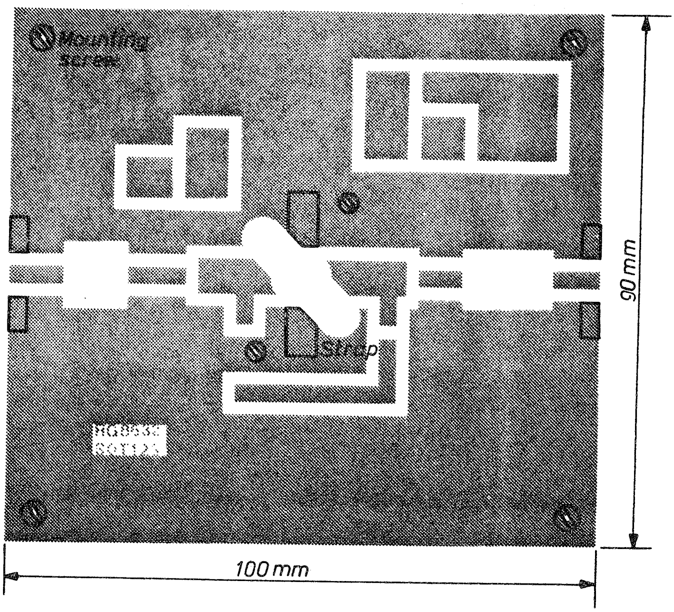


Fig.7 Printed circuit board

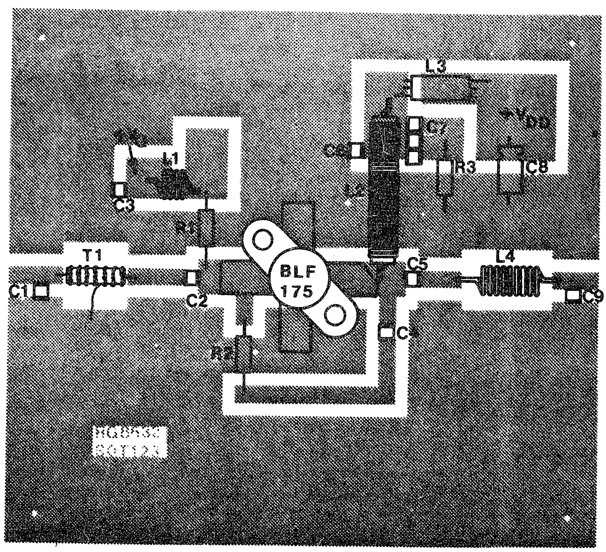


Fig.8 Component layout

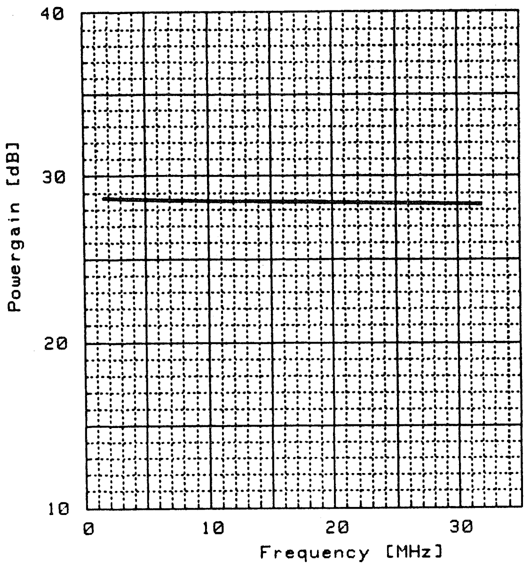


Fig.9 Powergain versus frequency

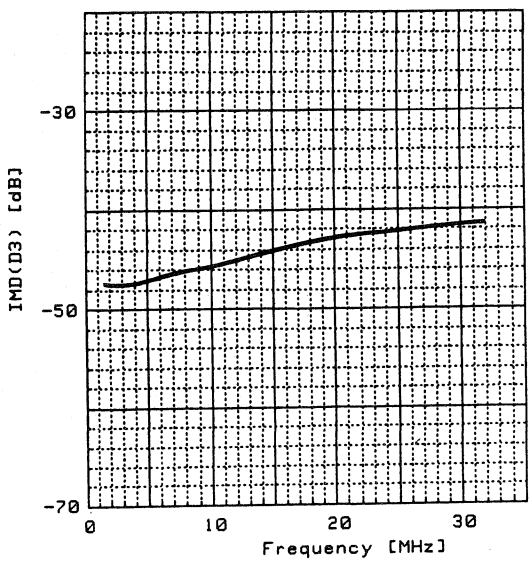


Fig.10 3rd order IMD versus frequency

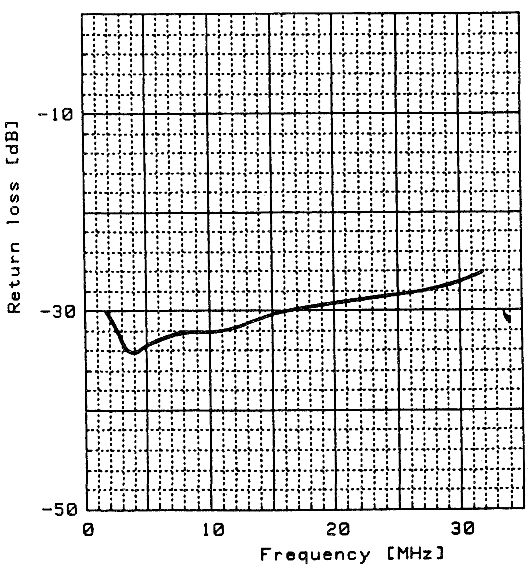


Fig.11 Input return loss versus frequency

Conditions: Vds = 50 Volt
Idq = 800 mA
Po = 8 W PEP
Th = 25 °C
Tone seperation = 1 kHz

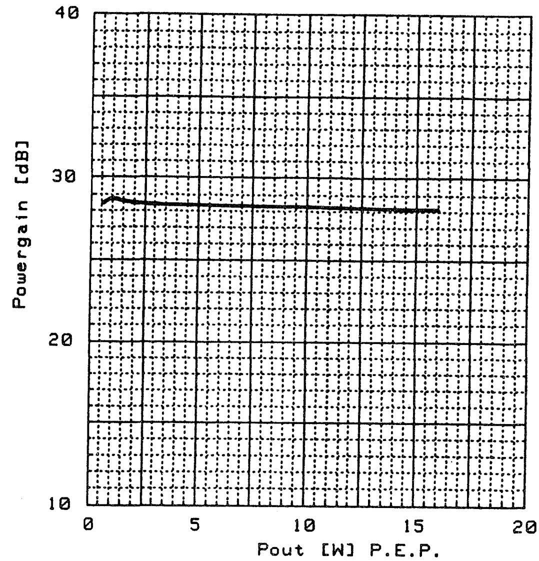


Fig. 12 Powergain versus output power

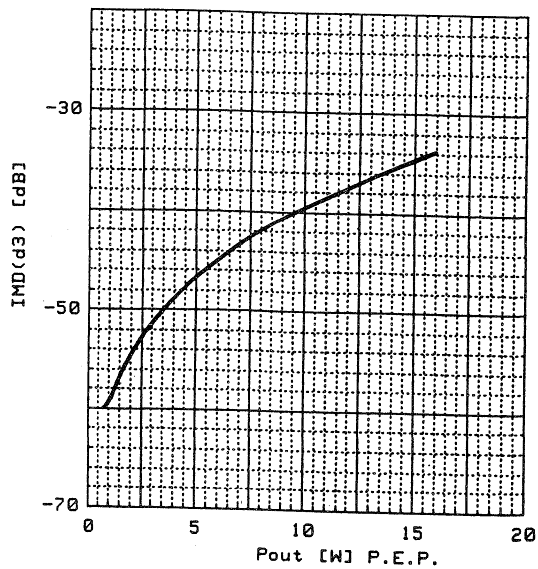


Fig. 13 3th order IMD versus output power

Conditions: Vds = 50 Volt
Idq = 800 mA
Th = 25 °C
f = 28 MHz (p-q=1kHz)

PRODUCT GROUP

PHILIPS SPECIALTIES AND DIODES APPLICATION

NIJMEGEN

Report nr. : NCO 8801
Author : A.H.Hilbers
Date : June 16, 1988

THE BLF246 AS AN H.F.-S.S.B. AMPLIFIER

SUMMARY

This report gives information on the BLF246 as a linear amplifier at 28 MHz for S.S.B. signals.

Typically the device produces an output power of 80 W P.E.P. at an I.M. distortion of -34 dB.
At a supply voltage of 28 V the power gain is 20 dB and the 2-tone efficiency 40 %.

For the design of wideband amplifiers in the range of 1.5 - 30 MHz additional information is presented showing that the power gain drops to appr. 19 dB with a variation of +/- 0.4 dB.

THE BLF246 AS AN H.F.-S.S.B. AMPLIFIER

1. Introduction

The BLF246 is an R.F. power MOS-transistor in SOT-121 package specified at a frequency of 108 MHz and a supply voltage of 28 V for an output power of 80 W C.W.

During the development period this device has also been tested in a class-AB amplifier at 28 MHz to investigate its behaviour as a linear amplifier for possible application in the H.F. band (1.5 - 30 MHz) with S.S.B. modulation.
In that case the 3rd and 5th order intermodulation products are of special interest.

2. Narrow band test at 28 MHz

The R.F. circuit used for this purpose is depicted in Fig. 1.
The optimum drain quiescent current for linear operation of this device in class-AB is 0.6 A.
This is achieved with a Vgs which is appr. 0.48 V higher than the specified Vgs(th).

The output circuit has been aligned with a dummy load consisting of the parallel connection of a 3.5 Ohms resistor and a 400 pF capacitor. This results in a load seen by the transistor of appr. 3.3 Ohms with a negligible reactive part.

The capacitors C9 and C10 are used to reduce the second harmonic voltage at the drain.
In a wideband amplifier this is not necessary because the same effect is obtained by other means like a centre-tapped drain choke in a push-pull amplifier.

The input circuit is aligned for minimum reflection at 50 % of the maximum output power.

The gate-source damping resistors R1 and R2 are needed for 2 reasons:

a. stability i.e. to prevent oscillation when the output circuit is detuned.
b. low intermodulation distortion. In general it can be said that I.M. distortion is improved by lowering the value of these resistors. Of course they will influence the power gain of the amplifier.

The average performance of the BLF246 in this amplifier is shown in Fig. 2 through 5:
Fig. 2: power gain versus O.P. power.
Fig. 3: 2-tone efficiency versus O.P. power.
Fig. 4: O.P. power versus drive power.
Fig. 5: d3 and d5 versus O.P. power.

At an output power of 80 W P.E.P. the average performance is:

Gp = 20 dB	d3 = -34 dB
Eff. = 40 % (2-tone)	d5 = -42 dB

The distortion products have been measured with respect to one tone.

It must be mentioned that the information given here is typical i.e. it is not guaranteed by measurements on individual transistors.

3. Wideband operation

For operation in the H.F. range (1.5 - 30 MHz) it will be necessary to reduce the value of Rgs from 18 Ohms as used in the narrow band amplifier down to 12 Ohms to obtain a more constant power gain over the band and to achieve a smaller variation in the input impedance to allow an easier matching to a driver stage or a 50 Ohms source.

Fig. 6 shows the results of a computer calculation on this basis. It appears that the power gain is reduced by appr. 1 dB while the gain variation is then +/- 0.4 dB.

The optimum load impedance is still appr. 3.3 Ohm with a negligible reactive part.

It must be mentioned that the figures given for power gain and input impedance include already the effects of an Rgs of 12 Ohms.

As mentioned earlier the lowering of Rgs has a positive effect on the intermodulation behaviour.

For the design of a suitable wideband input matching network one is referred to application report nr. RNR-1-259-1987.

4. Acknowledgement

The 28 MHz amplifier described in this report was designed and constructed by Mr.G.Lukkassen of our laboratory. He also made the various measurements.

A.H.Hilbers

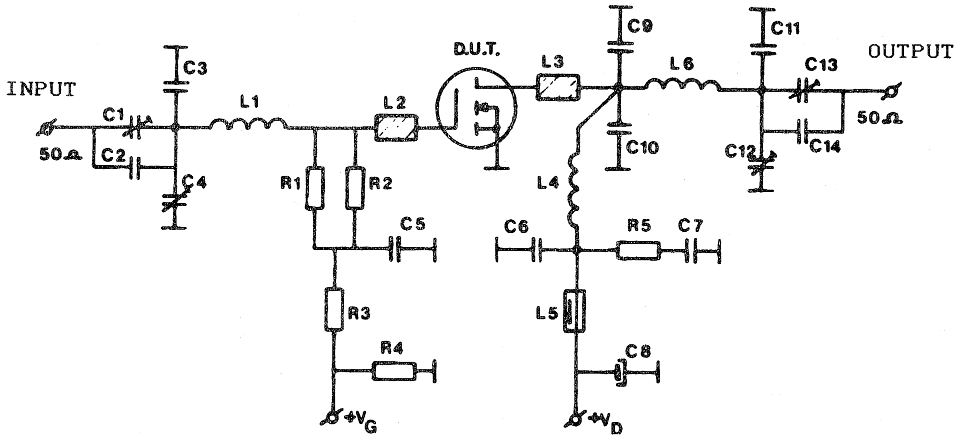


Fig. 1: BLF246 Testcircuit at 28 MHz

C1=C4= 6-80pF film dielectric trimmer (cat.nr.2222 809 07013)
C2=62pF multilayer ceramic chip capacitor *
C3=150pF multilayer ceramic chip capacitor *
C5=C6=100nF multilayer ceramic chip capacitor (cat.nr. 2222 852 47104)
C7=3x100nF multilayer ceramic chip capacitor (cat.nr. 2222 852 47104)
C8= 2.2μF electrolytic capacitor
C9=C10=C14=75pF multilayer ceramic chip capacitor *
C11=100pF multilayer ceramic chip capacitor *
C12=C13=7-100pF film dielectric trimmer (cat.nr.2222 809 07015)

L1=184nH,6 turns enamelled Cu-wire (0.7mm),int.dia:6mm
L2=L3=41.1Ω stripline (10mm x 6mm)
L4=278nH, 7 turns enamelled Cu-wire (1.5mm),int.dia: 8mm
L5= Ferroxcube h.f. choke, grade 3B (cat.nr.4312 020 36642)
L6= 131nH,4 turns enamelled Cu-wire (1.5mm), int.dia: 8mm

R1=R2=34.8Ω metal film resistor (cat.nr. 2322 153 53489)
R3= 1KΩ metal film resistor (cat.nr. 2322 151 71002)
R4= 1MΩ metal film resistor (cat.nr. 2322 151 71005)
R5= 10Ω metal film resistor (cat.nr. 2322 153 51009)

PC-board: double Cu-clad, 1.6mm PTFE fibre-glas dielectric (ε_r=2.2)

* American Technical Ceramics type 100B or capacitor of same quality

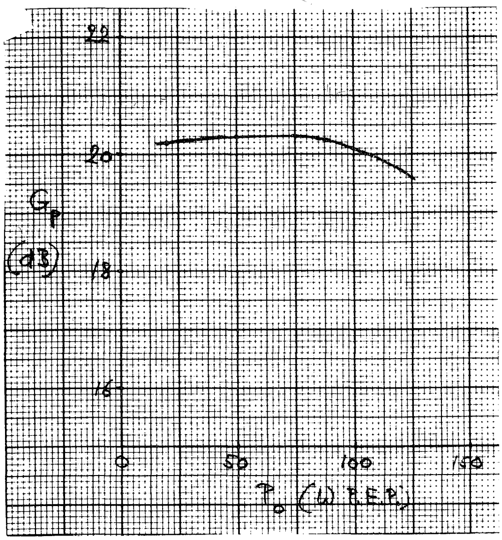


Fig. 2: Power gain versus output power

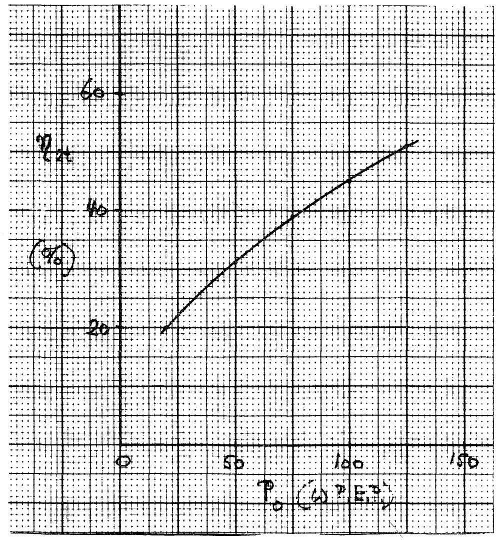


Fig. 3: 2-tone efficiency versus output power

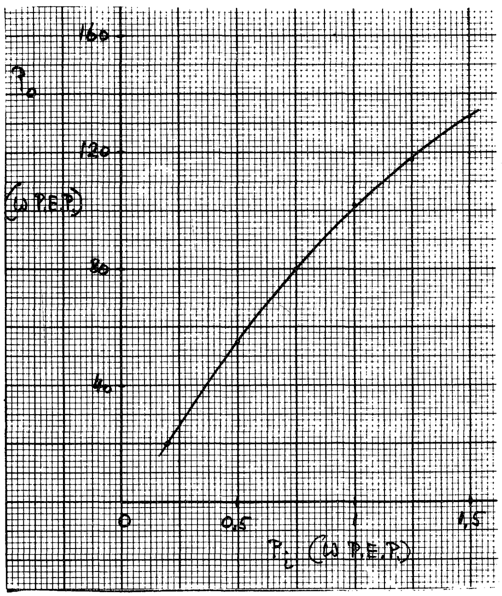


Fig. 4: Output power versus drive power

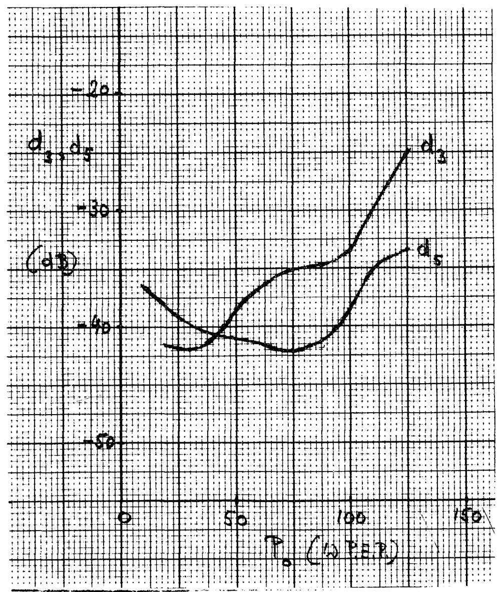


Fig. 5: Intermodulation distortion versus output power

Conditions:	f = 28 MHz	Idq = 0.6 A	Th = 25' C
	Vds = 28 V	Rgs = 18 Ohm	Rth mb-h = 0.2 K/W

BLF246 Vds= 28 V Po= 80 W P.E.P. Class-AB

f	*	G	*	Inp.Imp.	*	Load Imp.	*	Phi
MHz	*	dB	*	Ohm	*	Ohm	*	Deg
1.5	*	19.57	*	11.97 - j 0.51	*	3.34 + j 0.01	*	0.0
5.0	*	19.54	*	11.71 - j 1.67	*	3.34 + j 0.02	*	0.0
10.0	*	19.47	*	10.92 - j 3.06	*	3.33 + j 0.04	*	0.0
15.0	*	19.35	*	9.86 - j 4.03	*	3.32 + j 0.07	*	0.0
20.0	*	19.20	*	8.75 - j 4.59	*	3.31 + j 0.09	*	0.0
25.0	*	19.00	*	7.72 - j 4.84	*	3.30 + j 0.11	*	0.0
30.0	*	18.77	*	6.83 - j 4.86	*	3.28 + j 0.13	*	0.0

Fig. 6: Power gain, input and load impedance versus frequency.
Idq = 0.6 A; Rgs = 12 Ohm

NCO 8801: 6/6

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laboratory report

number : EC07806 date : 05-01-1979

title : WIDE-BAND LINEAR POWER AMPLIFIERS (470-860 MHz) WITH THE TRANSISTORS BLW32 AND BLW33

author : A.H. Hilbers and M.J. Köppen

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laboratory report

central application laboratory CAB
eindhoven - the netherlands

number:	ECO 7806	date:	05.01.1979
project:	6823	pages:	S6 + R42

title: WIDE-BAND LINEAR POWER AMPLIFIERS (470-860 MHz) WITH THE TRANSISTORS BLW32 AND BLW33

authors: A.H. Hilbers and M.J. Köppen

ABSTRACT:

For application in T.V. transposers for band IV/V (470-860 MHz) some wideband linear power amplifiers have been made with the BLW32 and BLW33. A single stage amplifier with BLW32 gave a minimum output power of 380mW peak sync. at a 2-tone i.m.d. of -47dB. The power gain was 11,4 ± 0,55dB. A similar amplifier with the BLW33 produced a minimum output of 790mW peak sync. at the same i.m.d. and a power gain of 10,4 ± 0,85dB. A 2-stage amplifier with the BLW32 as driver and BLW33 in the output showed a minimum output power of 750mW peak sync. at -47dB i.m.d. with an overall power gain of 22,5 ± 2dB.

Appr. R.A. Pölzl

Advies Octrooi d.d. 11 jan. 1979	X AV	GV		B		BL
Opgave Mamo d.d. 17 jan. 1979	X AV	X GV	X SP	B		BL
Datum: -5 jan. 1979	Mamo					

SUMMARY

Figs. A and B show the circuit diagrams of wide-band amplifiers with BLW32 and BLW33 resp.

These amplifiers operate in class A. The d.c. adjustment is controlled by means of bias networks with BD136.
The typical operation points are:

BLW32: V_{CE} = 22,5V and I_c = 165mA.
BLW33: V_{CE} = 22,5V and I_c = 330mA.

Both circuits are set-up for T.V. transposer service.
They cover the bands IV-V (470-860MHz) completely.

The circuits are printed on double copper clad p.c. board.
To keep the losses in the transforming strip transmission lines sufficiently low the dielectric is P.T.F.E. glass-fibre with ε_r = 2,74 having a thickness of 1/16 inch.

The fixed capacitors of the r.f. chain are of the multilayer ceramic chip type, whilst the variable capacitors have P.T.F.E. isolation or they are of the micro thin-trim type.

All typical elements to assure stable and spurious-free operation are applied. These decoupling elements have been chosen so that the effect on the video step response is acceptably small.

The designs have been optimized with the aid of a computer what means that the optimum collector load for the BLW32, varying typically between 41,5 + j39,4 ohms at 470MHz and 17,7 + j33,0 ohms at 860 MHz is transformed to the 50 ohms output terminal impedance with the aid of a Chebyshev bandpass filter (Ref. 3).

The same has been arranged for the BLW33, in which the typical load impedance varies between 27,1 + j17,4 ohms at 470MHz and 14,5 + j17,3 ohms at 860 MHz.

On the input side, the impedance for the BLW32 varies from 4,59 + j1,67 ohms at 470 MHz to 4,12 + j5,86 ohms at 860 MHz. These figures are for the BLW33 resp. 3,60 + j1,90 ohms and 3,19 + j4,90 ohms.

The calculated gain variation in the mentioned frequency band is 16,5 to 11,4dB for the BLW32 and 15,0 to 10,1dB for the BLW33. This gain variation has been reduced by permitting input mismatch at lower frequencies.
This has been done according to a method being described in Ref. 4. The calculated values derived from this method are optimized with a computer.

The typical results obtained after practical optimization with a network analyzer are summarized below:

BLW32 S21 (power gain) = 11,4 $\pm$ 0,55dB
S11 = -2,4 to -20,7dB
S22 $\leq$ -8dB

BLW33 S21 = 10,4 $\pm$ 0,85dB
S11 = -2 to -20,6dB
S22 $\leq$ -6dB

Both amplifiers have a high degree of linearity.
The output power versus frequency for a two-tone linearity of $d_3 \leq$ -47dB is shown in Figs. C and D.

As an experiment, both single amplifiers have been cascaded and combined on one p.c. board. Fig. F shows the complete circuit diagram. It has the consequence that the driver has to accept the mismatch caused by the final stage. In spite of this drawback the results were satisfying. The total gain (S21) became 22,5 $\pm$ 2dB, whilst the S11 and S22 still have the same trend as for the single ones. The minimum power for -47dB two-tone i.m.d. became 0,75 Watt (see Fig. E).

Having completed the amplifiers three-tone tests (-7, -8 and -16dB) were done at three channels (21, 39 and 70) for an i.m.d. of -60, -56 and -52dB.

These channels correspond with carrier frequencies of resp. 471,25, 615,25 and 863,25MHz.

It proved that there is a reasonable correlation between the two- and three-tone results.

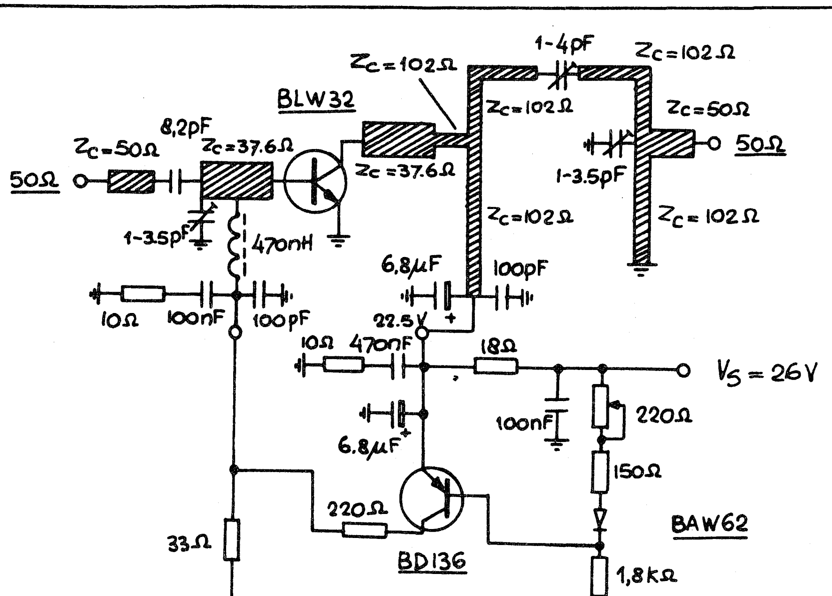

Fig. A

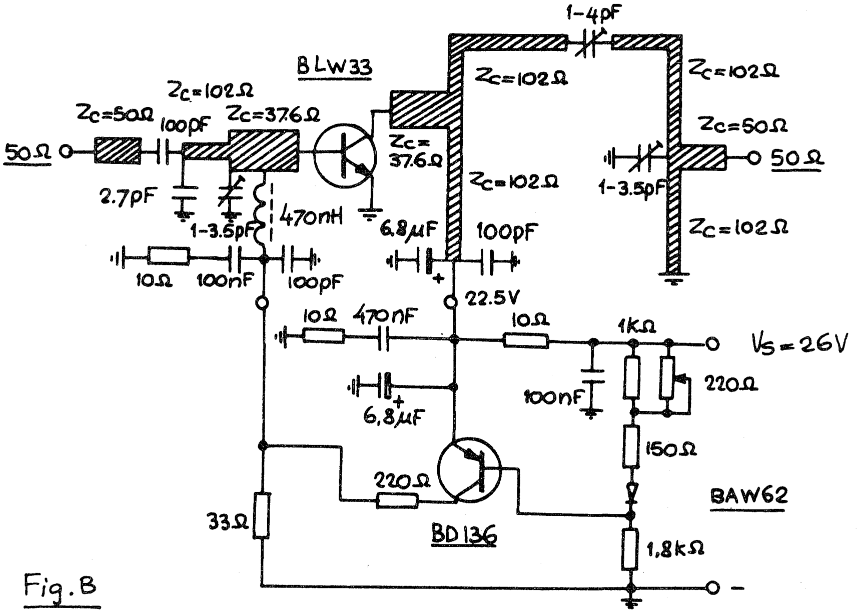

Fig. B

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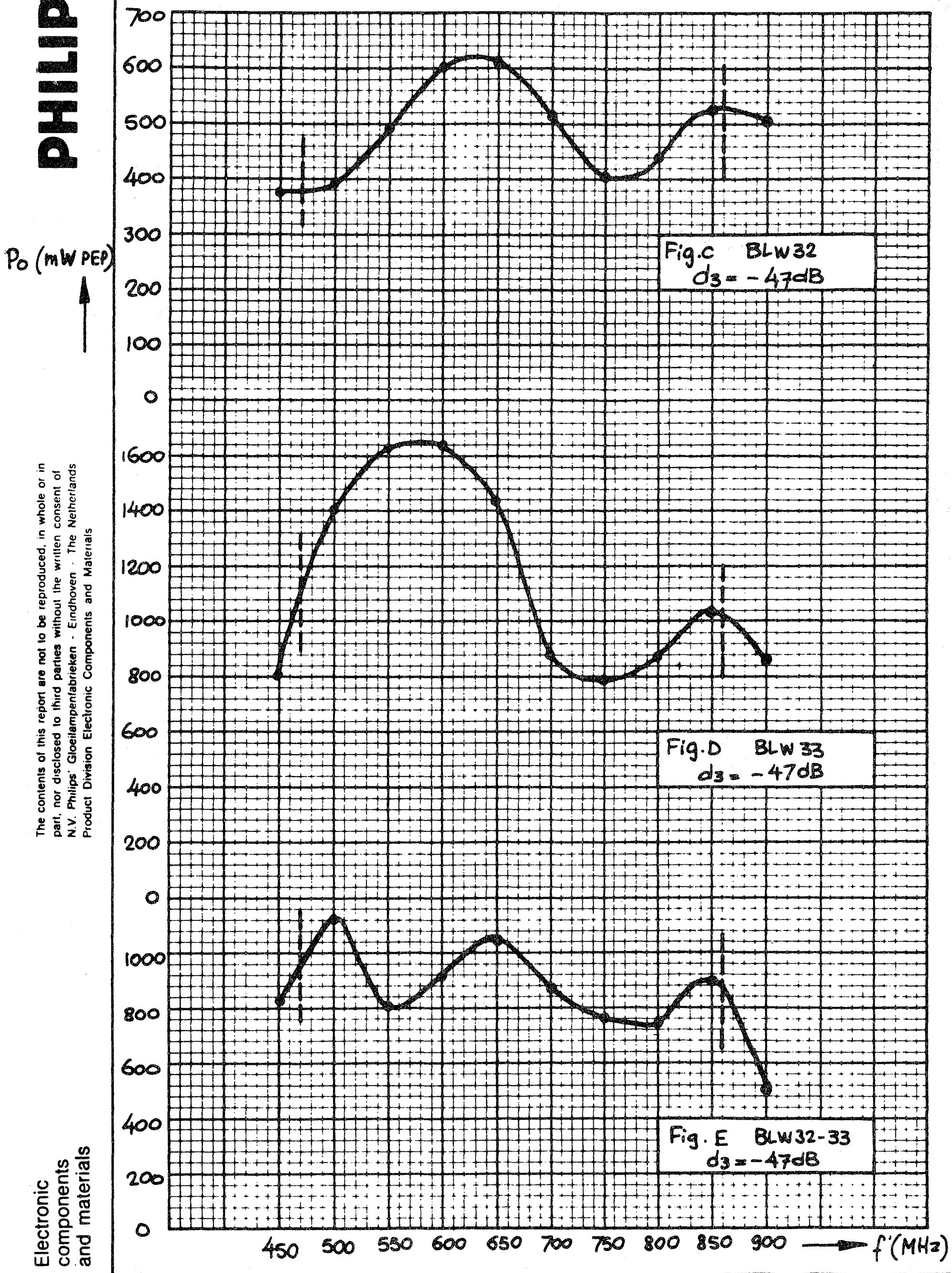

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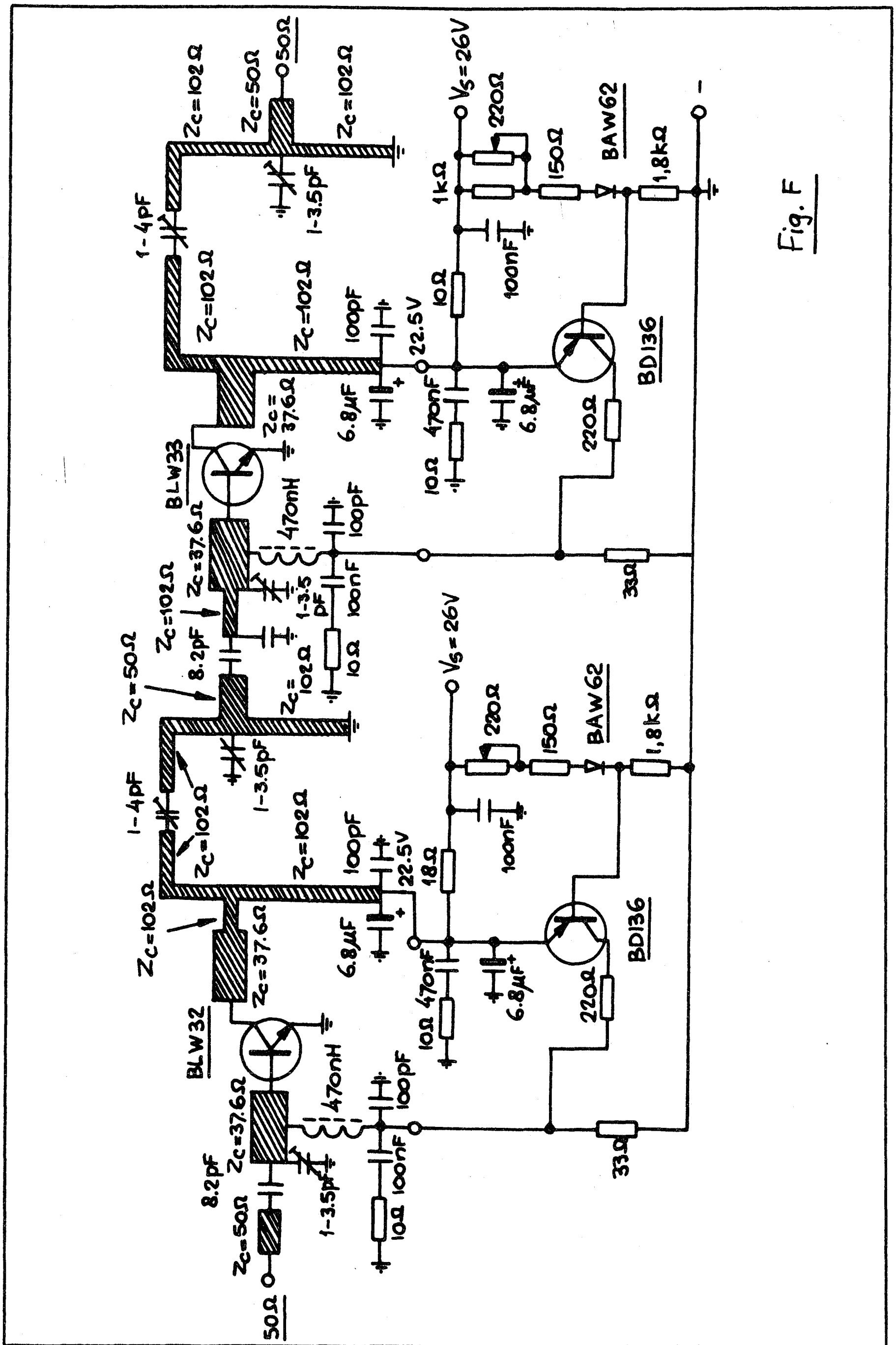

Fig. F

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1. INTRODUCTION

This report describes the theoretical aspects and practical realisation of some wide-band UHF power amplifiers for TV transposer service in bands IV and V (470-860 MHz).

The amplifiers are designed with the new BLW32 and BLW33 transistors. These devices are successors of the BLX96 and BLX97 for resp. 0,5 and 1 Watt peak sync output and are also developed for ultra linear applications. For this purpose the transistors have to operate in class A.

The power gain of both types is typically 5dB more than for the original devices. Because of this higher gain they are excellently suited for the realisation of wide-band type amplifiers.

The transistors are encapsulated in a ¼ inch capstan envelope with ceramic cap.

2. THEORETICAL CONSIDERATIONS

2.1. The equivalent circuit of the transistor input and output of the BLW32 and BLW33

For class A operation the BLW32 and BLW33 are specified as follows:

BLW32: V_{CE} = 25V and I_C = 150mA

BLW33: V_{CE} = 25V and I_C = 300mA

Although the power gain, the input impedance and the optimum narrow band load impedance versus frequency are given in the Data sheets for the above operating points, the actual class A operation points chosen differ somewhat for the circuits described in this report.

Namely, from earlier investigations it is known (see for instance report ECO 7704, Ref. 5) that the wide-band properties are more favourable for lower load impedances. So, we have chosen for the following class A operation points:

BLW32: V_{CE} = 22,5V and I_c = 165mA
BLW33: V_{CE} = 22,5V and I_c = 330mA

The corresponding typical gain, input and load impedances have been calculated (Ref. 1, 2). The values thus obtained for three frequencies are given below:

BLW32:

f (MHz)	gain (dB)	R_i (series) (Ohm)	X_i (series) (Ohm)	R_L (series) (Ohm)	X_L (series) (Ohm)
470	16,5	4,59	1,67	41,5	39,4
636	14,0	4,42	3,61	28,9	38,0
860	11,4	4,12	5,86	17,7	33,0

BLW33:

f (MHz)	gain (dB)	R_i (series) (Ohm)	X_i (series) (Ohm)	R_L (series) (Ohm)	X_L (series) (Ohm)
470	15,0	3,60	1,90	27,1	17,4
636	12,5	3,45	3,26	21,0	18,4
860	10,1	3,19	4,90	14,5	17,3

To facilitate calculations approximate equivalent circuits for the transistor input and output impedances can be given. They are shown in Figs. 1 and 2.

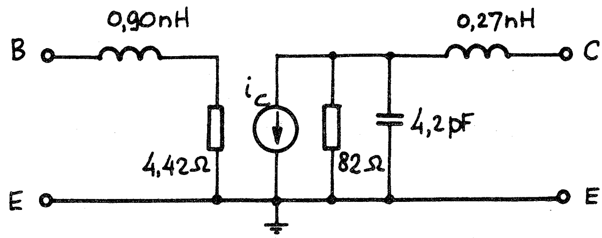

Fig. 1.: BLW32

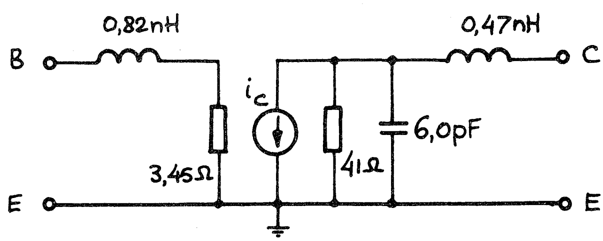

Fig. 2.: BLW33

2.2. The output networks

The circuits will be designed on printed circuit boards with PTFE-fibre glass as a dielectric with an ε_r = 2,74 and a thickness of 1/16 inch.

The input and output network start with a piece of stripline having a width of 6 mm, being the width of the base and collector leads. For a dielectric of 1/16 inch the characteristic impedance Z_c is 37,6 ohms. The length for the collector leads amounts to 3 mm. The base leads are different in length.

As the output impedance of both transistors is rather high a Chebyshev bandpass-filter configuration will be chosen to match it to the 50 ohms load. As will be shown later 6 elements are sufficient to obtain an acceptable VSWR through the band.

The bandpass filter is derived from a low-pass prototype having a cut-off frequency equal to the bandwidth of the final filter, i.e. 860-470=390 MHz and a characteristic resistance of 50 ohms (see Ref. 1). The transformation procedure is depicted in Figs. 3-7. These figures are for the BLW32.

To determine $C_1 = C_3$ in Fig. 3 we have to keep in mind that the input (and output) RC-product remains constant with impedance transformation, so: $C_1 = C_3 = 82 \times 4{,}2/50 = 6{,}9\text{pF}$.

Next, we determine the quantity γ being equal to :

$$\gamma = 1/(\omega_c . C_1 . R)$$

in which ω_c is the cut-off frequency of the low-pass prototype and $C_1.R$ is the already mentioned time constant, so:

$$\gamma = 1/(2.\pi.390.10^6.6{,}9.10^{-12}.50) = 1{,}183$$

Then L_2 can be calculated according to:

$$L_2 = \{R/\omega_c\}\{2\gamma/(\gamma^2 + 0{,}75)\}$$

in which R is the characteristic resistance of the filter being 50 ohms, so:

$$L_2 = \{50/(2.\pi.390.10^6)\}\{2.1{,}183/(1{,}399+0{,}75)\} = 22{,}46\text{nH}$$

The transformation from low-pass to bandpass is made by:

a shunting each capacitor by an inductor and

b putting a capacitor in series with each inductor, such that resonance is obtained at the geometric mean frequency of the band:

$$f_0 = \sqrt{860.470} = 635{,}8 \text{ MHz}.$$

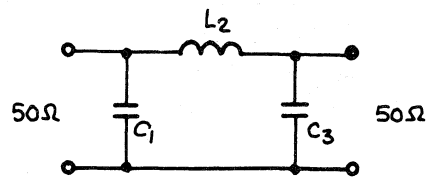

Fig.3

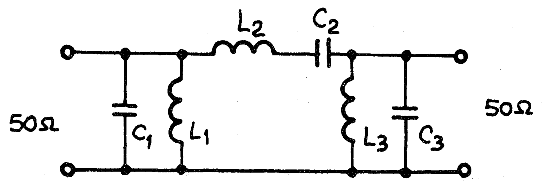

Fig.4

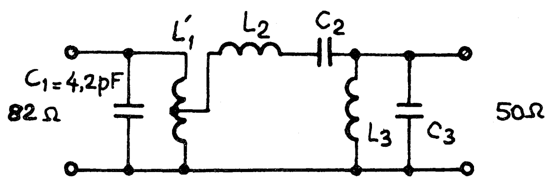

Fig.5

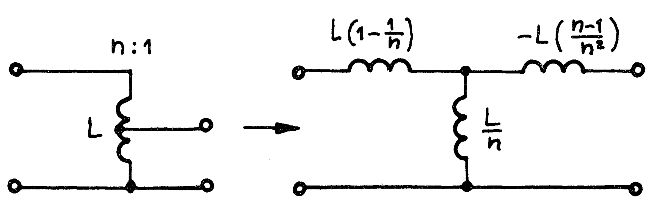

Fig.6

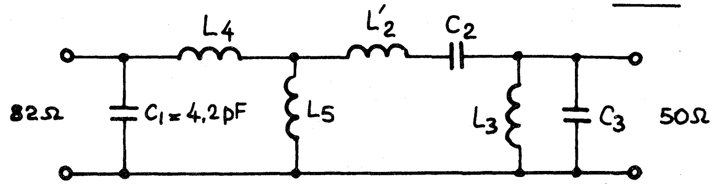

Fig.7

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This means in Fig. 4 that:

$L_1 = L_3 = 9{,}081$ nH
$C_2 = 2{,}79$ pF

Now we transform the input part of the filter to the actual output impedance of the transistor (see Fig. 5). This is done with an ideal transformer having a primary inductance, $L_1' = 82/50.9{,}081 = 14{,}89$nH and a transformation ratio n:1 in which $n = \sqrt{82/50} = 1{,}281$.

This ideal transformer is subjected to a Norton transformation (see Fig. 6) in which the inductances become 3,27 nH, 11,62 nH and -2,55 nH resp. The final situation is then depicted in Fig. 7 with $L_4 = 3{,}27$ nH, $L_5 = 11{,}62$ nH and $L_2' = L_2 - 2{,}55$ nH $= 22{,}46 - 2{,}55 = 19{,}91$ nH.

The resulting maximum VSWR is equal to:

$S = \{(x^3+1)/(x^3-1)\}^2$ in which:

$x = \gamma + \sqrt{\gamma^2 + 1}$

For our filter this becomes: S = 1,22, which is a very acceptable value justifying the conclusion that the number of filter elements is sufficient.
The next step is the transformation of this filter to a strip-line circuit. This has been done in the same way as described on pages R6 to R8 of Ref. 5. If a transmission line is shorter than 1/8 of a wavelength the following equivalence is reasonably accurate (Fig. 8):

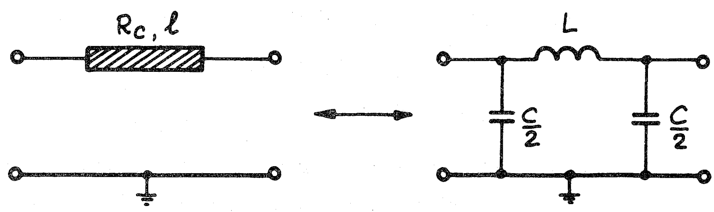

Fig. 8.

$$L = \frac{R_c \cdot l}{v} \quad \text{and} \quad C = \frac{1}{R_c \cdot v} ,$$

in which v is the propagation speed being 3.10^8 m/s for air-line. Starting from Fig. 7 we must keep in mind that C_1 and part of L_4 (viz. 0,27nH) are inside the transistor.
The inductors are then replaced by striplines. The consequence of this transformation is that at some points in the circuit parasitic capacitances are introduced.
This can be corrected by adjusting the values of L_5 and C_3.
A second reason for correcting the latter is its parasitic series inductance. The remaining part of L_4 is composed of 2 pieces of stripline with different characteristic impedances. The part with the lower R_c serves to accomodate the collector lead of the transistor. C_2 has a parasitic series inductance of appr. 1nH which must be subtracted from L_2'.
The remaining part is split up in 2 equal pieces on both sides of C_2. The result of the transformation described above is shown in Fig. 9 and table 1.

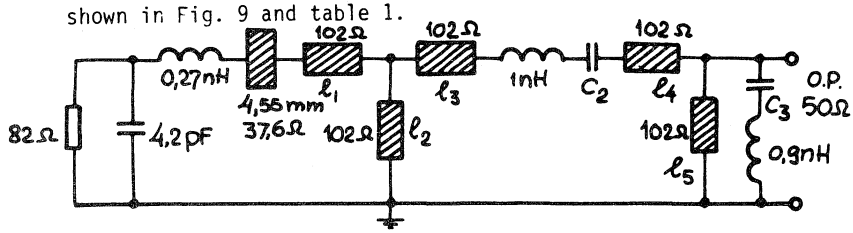

Fig. 9. See also table 1.

An exact analysis of this circuit results in a maximum output VSWR of 1,66 which is rather high. Therefore the complete circuit is subjected to a computer optimization procedure with the object to reduce the maximum output VSWR to the lowest possible value. The results are given in table I.

Element	Before optim.	After optim.	Unit
l_1	7,15	6,84	mm
l_2	28,4	27,7	mm
l_3	27,8	29,8	mm
C_2	2,79	2,76	pF
l_4	27,8	29,8	mm
l_5	26,7	33,4	mm
C_3	5,14	4,01	pF
S_{max}	1,66	1,35	-

Table 1.

As a final step the practical dimensions of the striplines must be determined. The p.c. board material is PTFE fibre-glass with a thickness of 1/16 inch and a dielectric constant ε_r of 2,74. The lines with R_c = 102 ohms must have a width of 1 mm. The length reduction factor is 1,43. The collector pad (R_c = 37,6 ohms) must have a width of 6 mm and its length reduction factor is 1,52. The results of the latter transformation can be found in the final circuit diagram and parts list.

2.3. The input networks

The approach followed for these networks is rather similar to those described in Refs. 4 and 5. However, there is one exception:

Because of the higher input impedance of the BLW32, as compared to the BLW98, a smaller amount of transformation is required. As a consequence of this a single section matching network is sufficient. Generally speaking it can be said that this section must have a loaded Q-factor of appr. 4 to produce the required gain compensation.

This means that the total inductive reactance (of transistor and matching network) must be 4 times the input resistance of the transistor (see Fig. 10).

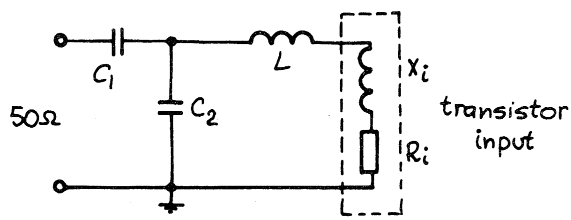

Fig. 10.

The tuning of this circuit is then at appr. 860 MHz.

The single section transforms the transistor impedance to a value above 50 ohms, so that some "back-transformation" is required. This is done by reducing the value of the blocking capacitor C_1 (8,2pF instead of 100pF).

In case of the BLW33 the input impedance is somewhat lower, so a 2 section network is advisable. Here, the blocking capacitor is 100pF.

The final steps are the same as in the previous section, viz. transformation to stripline and computer optimization.
The results can again be found in the circuit diagrams.

3. THE SINGLE AMPLIFIERS WITH BLW32 AND BLW33

3.1. Practical considerations

On previous pages the theoretical approach has been discussed.
In practice it was the intention to realize small compact amplifiers on a p.c. board with the input and output terminals (R_C = 50 ohms) in-line for easy cascading of several amplifiers.

The p.c. board needs to be double copper clad and has a PTFE fibre-glass dielectric for low losses at UHF. With a typical ε_r of 2.74 we had the choice between a thickness of 1/16 and 1/32 inch.

In principle 1/32 inch is possible but especially the 102 ohms lines in this design become so narrow, that the risk of under etching is too high.
So, we have decided to use a dielectric of 1/16 inch, what means that these 102 ohms lines are 1 mm wide.

Another problem being met with the application of 1/32 inch is the very small surface for soldering of the multi-layer chips and also an unusual ratio between the width of chip and track.

Fig. 11 shows the circuit diagram of the BLW32 class A amplifier with biasing, whilst in Fig. 12 the circuit is drawn for the BLW33. Both circuits contain a bias network with a PNP transistor type BD136.

The p.c. boards are resp. in Figs. 13 and 14 and the lay-out in Figs. 15 and 16.
For a correct earthing the upper earth sheet parts are directly connected to the lower sheet by soldering copper straps at the edges of the p.c. board.
Figs. 15 and 16 illustrate how it has been arranged.
The black parts are the soldered copper straps.

The emitters were grounded as short as possible by applying copper straps under the emitter leads. For that reason the holes in the board were square instead of round.

All components are situated on one side of the board viz. the side of the tracks.

The transistors were screwed to an extruded aluminium heatsink, resulting in an average stud temperature of appr. 50^{o}C.
In both circuits tuning capacitors are applied. They are of the film dielectric type with three tags. Both earth terminals are fed through holes and soldered to the upper and lower earth plane.

In the experimental phase, the coupling capacitors in the collector bandpass sections are of the "Micro thin-trim" type 1,0 - 4,0 pF. Later, they can probably be replaced by fixed multilayer chip capacitors.

The coaxial connectors are of the SMA 50 ohms type.
They are soldered to the upper and lower earth plane.

In general, earth connections have to be made as short as possible.

3.2. Practical optimization method

Both prototypes were built according to the results of the theoretical part. It proved that such a theoretical approach was very valuable. It led, via some relatively small modifications to the ultimate design.

Optimization on a small signal basis has been done with the circuits inserted in a network analyzer chain, having swept S (scattering)-parameter facilities. From this set-up one can examine under dynamical conditions the effect on the power gain (represented by S_{21}), input- and output impedances (resp. S_{11} and S_{22}) and feedback (S_{12}) by tuning and modifications on the circuit of the amplifier.

The correction of power gain over the whole 470-860 MHz range is arranged by applying appropriate mismatch. The resulting high VSWR, especially at low frequencies, is not reduced in an equalizer.

So, tuning on an S-parameter set-up means that the S_{21} has to be optimized for flat gain, and at the same time the input reflection damping S_{11} tuned for minimum at the highest frequency of 860 MHz.

3.3. Measured results

Figs. 17a, b and c and 18a, b and c show the practical results for resp. the BLW32 and BLW33 amplifiers.
The results for two units are both given. They are numbered (1) and (2).

For a clear interpretation of the S_{11} and S_{22} readings the expression for the reflection damping and some figures are given below:

$$\text{refl. damping} = 20 \log (S+1)/(S-1) \text{ dB}$$

in which S is the voltage standing wave ratio (VSWR).

refl. damping (dB)	refl. coeff.	S
0	1	∞
4	0,631	4,42
8	0,398	2,32
12	0,251	1,67
16	0,158	1,38
20	0,1	1,22
24	0,063	1,13
28	0,04	1,08

Examining Figs. 17a and 18a one will see that in both cases, the reflection damping is about 20dB so the input VSWR is about 1,2 near 860 MHz, whilst the output VSWR, with respect to 50 ohms, varies between 3 and 1,06 over the band.

After optimization at small signals, the i.m.d. has been measured in a two-tone set-up. Although it is advised to apply the three-tone test method for determining the i.m.d. in case of TV systems, the two-tone test shows a good correlation with the former (Ref. 6).

In practice this means that for a three-tone test with the tones at -7, -8 and -16dB below the 0dB peak sync. level, the in-band i.m. product has to be at least -60dB down, whilst the same amplfier has to show a distance of two i.m. products d_3 of at least -47dB with respect to one of the equal tones in case of the two-tone test with equal peak output power.

The results are given in the curves of Figs. 19a, b and 20a, b.

For these amplifiers, of which two each have been constructed and measured, the results are according to the expected ones for wide-band operation.
At three channels the peak sync power for a three-tone i.m.d. of -60, -56 and -52dB has been measured too.
Fig. 21 gives an impression of the average -60dB results of the BLW32 and BLW33.

Channel nr.	BLW32	BLW33
21 (471,25 MHz)	445mW	1286mW
39 (615,25 MHz)	671mW	1308mW
70 (863,25 MHz)	500mW	941mW
P_0 sync. for -60dB i.m.d. (average of two amplifiers)		

Fig. 21.

Tables 2 and 3 show the more complete test results.

Comparing the afore mentioned three-tone (-60dB) results with the two-tone (-47dB) figures it appears that:

As for the single amplifiers the typical low frequency decoupling elements are divided over the bias circuit and the amplifier circuit. This was done because in case of concentrating all low-frequency decoupling elements on the bias board, parasitic oscillations occurred.

To get some idea of the reproduceability of practical amplifiers, two equivalent units have been built. They are screwed to heatsinks having a thermal resistance of appr. 2^{0}C/W. So for an average ambient temperature of 25^{0}C the stud temperature rises to about 48^{0}C for a power dissipation of about 11,5 Watts.

For the single amplifiers described, the straight forward method of small signal tuning at the S-parameter equipment and afterwards the two-tone test was followed. Via this method reliable results could be obtained.
However in case of the double amplifier it appeared to be rather difficult to find the optimum between an acceptable flat gain curve and sufficient output power with low i.m.d.

Because there is a correlation between the single tone 1dB compression point and the i.m.d. of a linear amplifier, the following dynamic way of tuning was applied:
For the BLW33 compression starts around 2W output. The gain of the complete amplifier is appr. 20dB so appr. 20mW (+13dBm) drive power should be sufficient. This amount of swept power is available from the sweep oscillator HP8620C in combination with RF plug-in unit HP86222A. Because of internal ALC the swept output power level is very stable and almost unaffectable by the strong input VSWR variations of the BLW32 amplifier.

- for the BLW32, the channel 21 and 39 results are somewhat better (3-18%) when measured under three-tone conditions, whilst the channel 70 figures are about 6% lower
- in case of the BLW33, the three tone results are somewhat different for both tested units. As an average they are in between those of the two-tone test.
- The gain figures are about the same.

  For a description of the two-tone and three-tone test chain and the way of testing one is referred to Refs. 5 and 6.

4. COMBINED AMPLIFIER WITH BLW32 AND BLW33

4.1. Practical considerations

As an experiment both amplifiers, being tuned for optimum performance, were cascaded. In that case the output of the BLW32 has to accept the mismatch, specially at lower frequencies, caused by the BLW33 amplifier.

The gain (S_{21}) becomes appr. 20dB, whilst the input and output reflection damping of the combination show somewhat different values. However, the trend is the same.

From this experiment the idea came to integrate both amplifiers on one p.c. board. The easiest solution was the direct connection of the two 50 ohms striplines, maintaining the blocking capacitor C_1 (Fig. 12) and concentrating the elements C_{11} (Fig. 11) and C_2 (Fig. 12) in a new single tuning element.

Fig. 22 shows the total circuit diagram Fig. 23 the p.c. board and Fig. 24 the lay-out of the amplifier part.

To create an amplitude variation the sweep generator is amplitude modulated (AM input) with a square wave pulse generated with a PM5715 pulse generator.

The output of the amplifier under test is now again attenuated (about 20dB), detected and supplied to an oscilloscope being horizontally driven by means of the available sweep output of the HP8620C.

Fig. 25 shows the block diagram of the measuring set-up, and Figs. 26a, b and c screen pictures of the swept output power for different input levels (+7, +10 and +13dBm) in which compression is clearly shown.
From above mentioned experiments it appeared that the blocking capacitor, C_{12} in Fig. 22, better could be decreased to 8,2 pF.

4.2. Measured results

Being tuned for a good large signal behaviour the i.m.d. results of both amplifiers constructed are given in Figs. 27a and 27b. Figs. 28a and 28b resp. show the gain curves for an input power of 5mW (+7 dBm) and Figs. 29a, b and c the S-parameter results.

The peak sync. power for a three-tone i.m.d. of -60, -56 and -52dB has been measured at three channels. Fig. 30 shows the -60dB results for the combination nr. (2).

Channel nr.	BLW32-BLW33
21 (471,25 MHz)	1045 mW
39 (615,25 MHz)	895 mW
70 (863,25 MHz)	587 mW
P_o sync for -60dB i.m.d. (amplifier nr. 2)	

Fig. 30.

Table 4 shows the more complete test results.

Comparing the two-tone (-47dB) and three-tone (-60dB) results it appears that:

- the combination BLW32/33 shows appr. equal results for channels 21 and 39. However the channel 70 results are somewhat lower.
- the gain figures are about the same.

5. CONCLUSIONS

5.1. On preceding pages the theoretical and practical designs have been described of some wide-band (470-860 MHz) high quality linear amplifiers, utilizing the BLW32 and BLW33 transistors operating in class A.

The class A operation points, in these wide-band applications, differ somewhat from those being published in the Data sheets for narrow band applications. This led to a better wide-band loading, what again resulted in a better i.m.d. performance.

5.2. The flatness of the power gain versus frequency is obtained by applying an increasing amount of mismatch when the frequency becomes lower.

This can be seen as a disadvantage of the system, because the preceding driver stage has to accept that mismatch.
In the present combination, however, the driver is so large that no difficulties are experienced.

6. RECOMMENDATIONS

6.1. A flat power gain characteristic combined with a low input VSWR may be obtained by:

a. applying an equalizing network at the input.

b. a coaxial isolator or circulator at the input.
However, these devices have restricted bandwidths.

c. applying the method being followed in C.A.B.-report ECO 7704 (Ref. 5) in which two equal amplifiers (with BLW98) have been connected in parallel with the aid of two wide-band 3dB-90^{0} coaxial hybrids on a 50 ohms basis.
In that configuration the output power will be nearly doubled, whilst the input VSWR of the system becomes max. 1,2, a value mainly given by the properties of these devices.
The reflected power will be absorbed in the resistor matching the isolated port.

A suitable type of hybrid is the ultra-miniature 3dB-90^{0} coupler, model 10264-3 (range 0,5 - 1,0 GHz) or 1H 0264-3 (range 0,44 - 0,88 GHz) from Anaren Microwave Inc.

Compared with the rest of the components such an arrangement is rather expensive.

If one wishes to construct the couplers oneself it can be done with "wire-line" type BH 10 of Sage. The assembling, however, asks for special tools.

6.2. In this report a number of measuring methods are given. It could be considered to apply the swept compression test method with an external wide-band amplifier (no need for linearity) and the external ALC possibility for testing amplifiers asking for more drive power.

7. REFERENCES

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Ref. 2: J. Mulder - Large signal behaviour of r.f. power transistors.
Part 2. Computer program,
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Ref. 3: G.L. Matthaei, L. Young and E.M.T. Jones - Microwaves filters, impedance-matching networks, and coupling structures.
Mc.Graw-Hill Book Company, New York.

Ref. 4: O. Pitzalis, Jr. and R.A. Gilson - Tables of impedance matching networks with approximate prescribed attenuation versus frequency slopes.
IEEE Transactions on microwave theory and techniques, Vol. MTT-19, no. 4, April 1971, pp. 381-386.

Ref. 5: A.H. Hilbers and M.J. Köppen - Design of a wide-band linear power amplifier (470 - 860 MHz) with two transistors BLW98. C.A.B. report ECO 7704.

PHILIPS

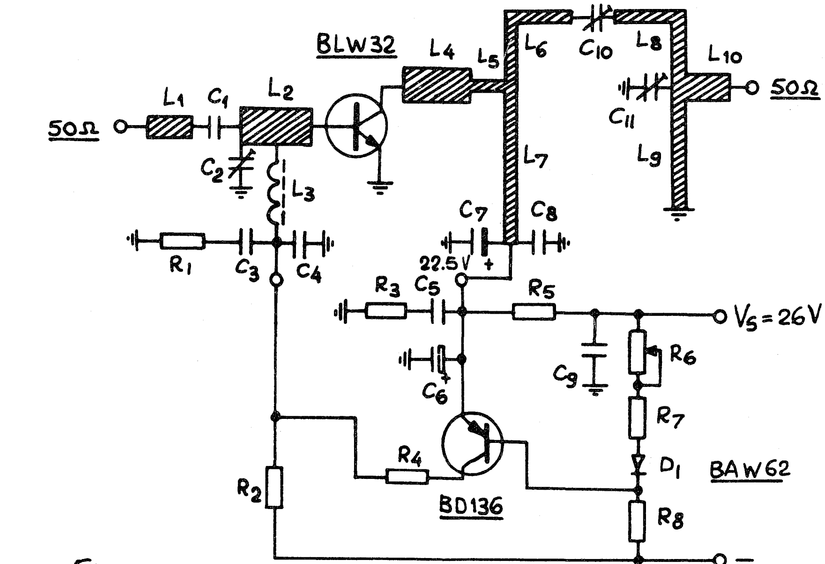

Fig. 11

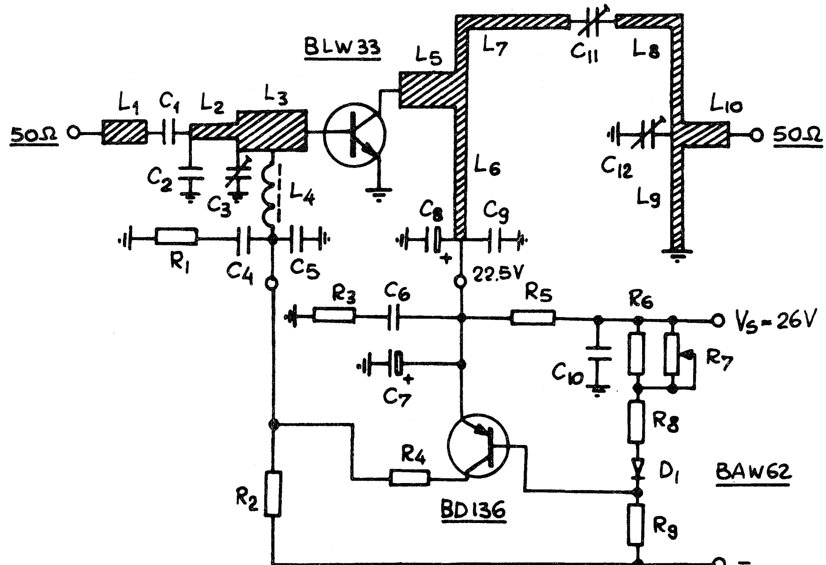

Fig. 12

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Ref. 6: M.J. Köppen - The BLX98 as a linear amplifier at 1 GHz.
C.A.B.-report ECO 7601.

8. LIST OF COMPONENTS

8.1. BLW32 amplifiers (Figs. 11, 13, 15)

C_1 = 8,2pF, multilayer ceramic chip capacitor, ATC (American Technical Ceramics) type 100A-8R2-J-Px-50.

$C_2 = C_{11}$ = 1 to 3,5 pF film dielectric trimmer (cat. no. 2222 809 05001).

$C_3 = C_9$ = 100nF, polyester capacitor

$C_4 = C_8$ = 100pF, multilayer ceramic chip capacitor (cat. no. 2222 852 13101).

C_5 = 470nF, polyester capacitor

$C_6 = C_7$ = 6,8 μF, 63V, electrolytic capacitor

C_{10} = 1 to 4pF, micro thin-trim, Tekelec Airtronic part no. AT9401-4-SL1.

$L_1 = L_{10}$ = stripline (Z_c = 50Ω), width 4,0 mm.

L_2 = stripline (Z_c = 37.6Ω), 11,5 x 6,0 mm^2.

L_3 = 470nH, microchoke

L_4 = stripline (Z_c = 37,6Ω), 3,0 x 6,0 mm^2.

L_5 = stripline (Z_c = 102Ω) , 4,8 x 1,0 mm^2.

$L_6 = L_8$ = stripline (Z_c = 102Ω), 20,8 x 1,0 mm^2.

L_7 = stripline (Z_c = 102Ω), 19,3 x 1,0 mm^2.

L_9 = stripline (Z_c = 102Ω), 23,4 x 1,0 mm^2.

R_1 = R_3 = 10Ω (± 5%) carbon resistor CR25 type.
R_2 = 33Ω (± 5%) carbon resistor CR25 type.
R_4 = 220Ω (± 5%) power metal film resistor PR37 type.
R_5 = 18Ω (± 5%) power metal film resistor PR52 type.
R_6 = 220Ω, cermet preset potentiometer.
R_7 = 150Ω (± 5%) carbon resistor CR25 type.
R_8 = 1,8kΩ (± 5%) carbon resistor CR25 type.

D_1 = BAW62.

8.2. BLW33 amplifier (Figs. 12, 14, 16)

C_1 = C_5 = C_9 = 100pF, multilayer ceramic chip capacitor (cat. no. 2222 852 13101).
C_2 = 2,7pF multilayer ceramic chip capacitor, ATC type 100A-2R7-B-Px-50.
C_3 = C_{12} = 1 to 3,5pF film dielectric trimmer (cat. no. 2222 809 05001).
C_4 = C_{10} = 100nF, polyester capacitor
C_6 = 470nF, polyester capacitor
C_7 = C_8 = 6,8 μF, 63V, electrolytic capacitor
C_{11} = 1 to 4 pF, micro thin-trim, Tekelec Airtronic part no. AT 9401-4-SL1.

L_1 = L_{10} = stripline (Z_c = 50Ω) width 4,0 mm.
L_2 = stripline (Z_c = 102Ω), 9,6 x 1,0 mm^2.
L_3 = stripline (Z_c = 37,6Ω), 5,3 x 6,0 mm^2.
L_4 = 470nH, microchoke
L_5 = stripline (Z_c = 37,6Ω), 3,0 x 6,0 mm^2.
L_6 = stripline (Z_c = 102Ω), 16,2 x 1,0 mm^2.
L_7 = L_8 = stripline (Z_c = 102Ω), 21,7 x 1,0 mm^2.
L_9 = stripline (Z_c = 102Ω), 20,4 x 1,0 mm^2.

$R_1 = R_3 = 10\Omega$ ($\pm$ 5%) carbon resistor CR25 type.
$R_2 = 33\Omega$ ($\pm$ 5%) carbon resistor CR25 type.
$R_4 = 220\Omega$ ($\pm$ 5%) power metal film resistor PR37 type.
$R_5 = 10\Omega$ ($\pm$ 5%) enamelled wire-wound resistor WR0617E style
$R_6 = 1k\Omega$ ($\pm$ 5%) carbon resistor CR25 style.
$R_7 = 220\Omega$, cermet preset potentiometer.
$R_8 = 150\Omega$ ($\pm$ 5%) carbon resistor CR25 style.
$R_9 = 1{,}8k\Omega$ ($\pm$ 5%) carbon resistor CR25 style.

D_1 = BAW62.

P.C. BOARDS
1/16 inch PTFE
double Cu clad

p.c. board
BLW32

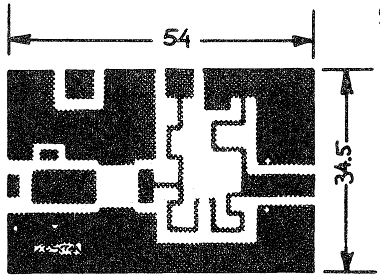


Fig. 13

p.c. board
BLW33

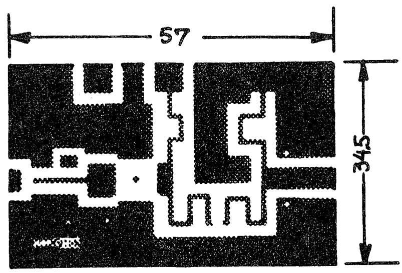
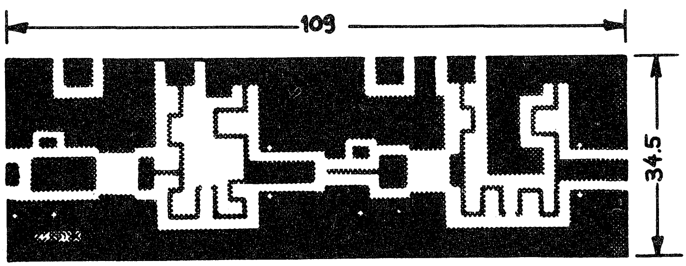


Fig. 14



p.c. board
BLW32-33

Fig. 23

LAY-OUT

1/16 inch PTFE double Cu clad

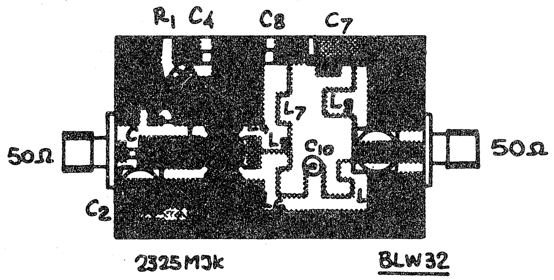


Fig. 15

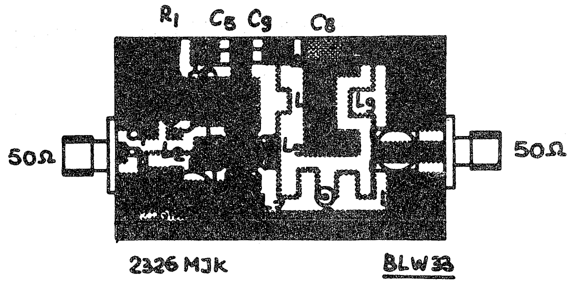


Fig. 16

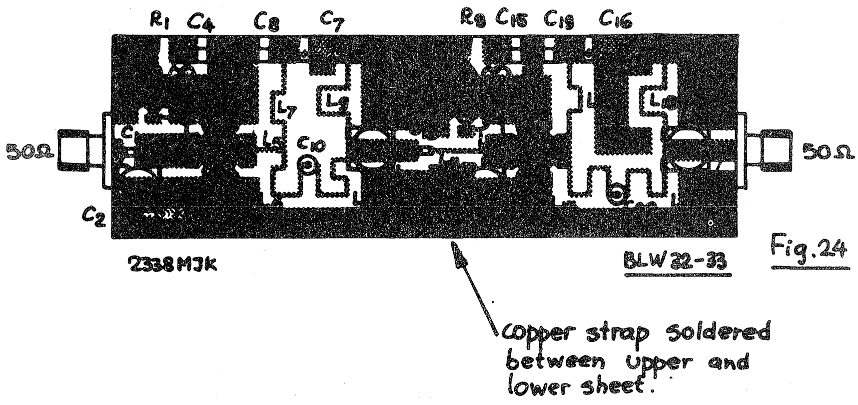


Fig. 24

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BLW32

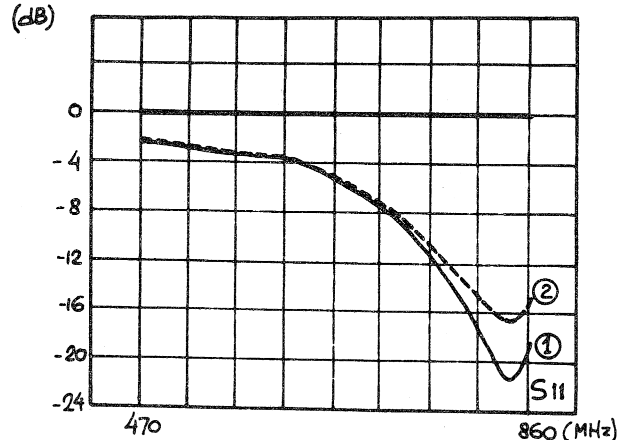

Fig. 17a

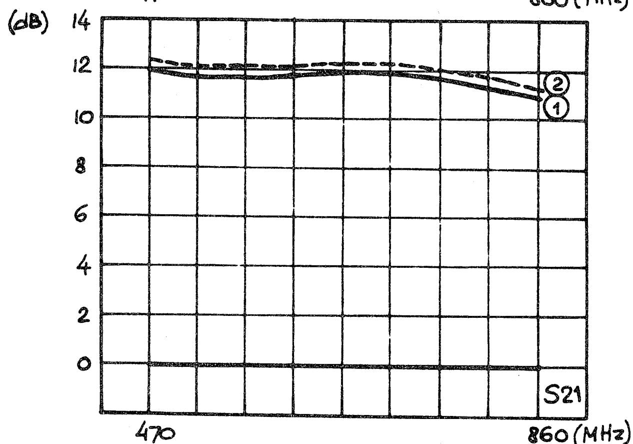

Fig. 17b

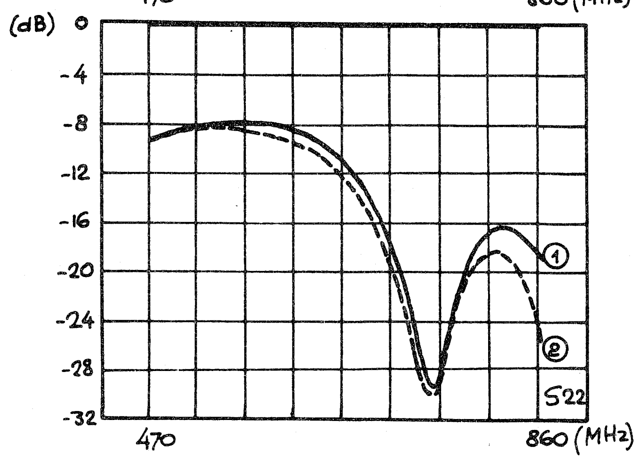

Fig. 17c

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BLW33

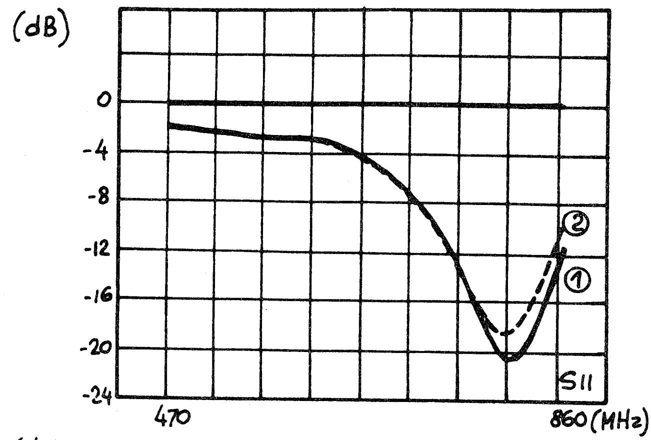

Fig. 18a

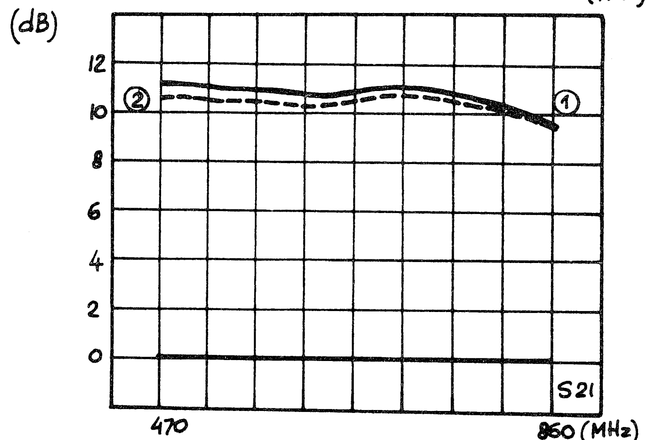

Fig. 18b

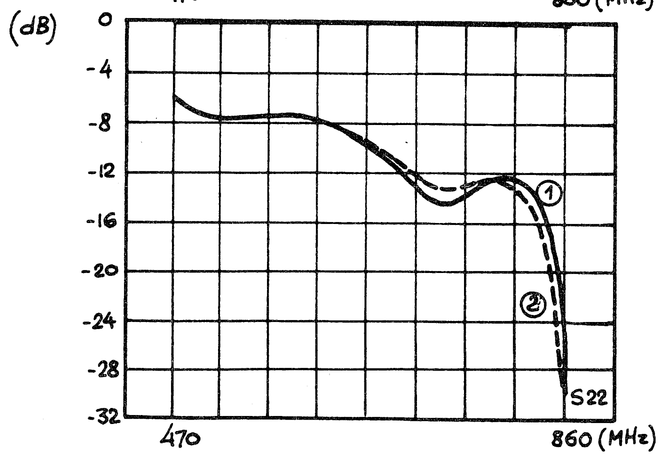

Fig. 18c

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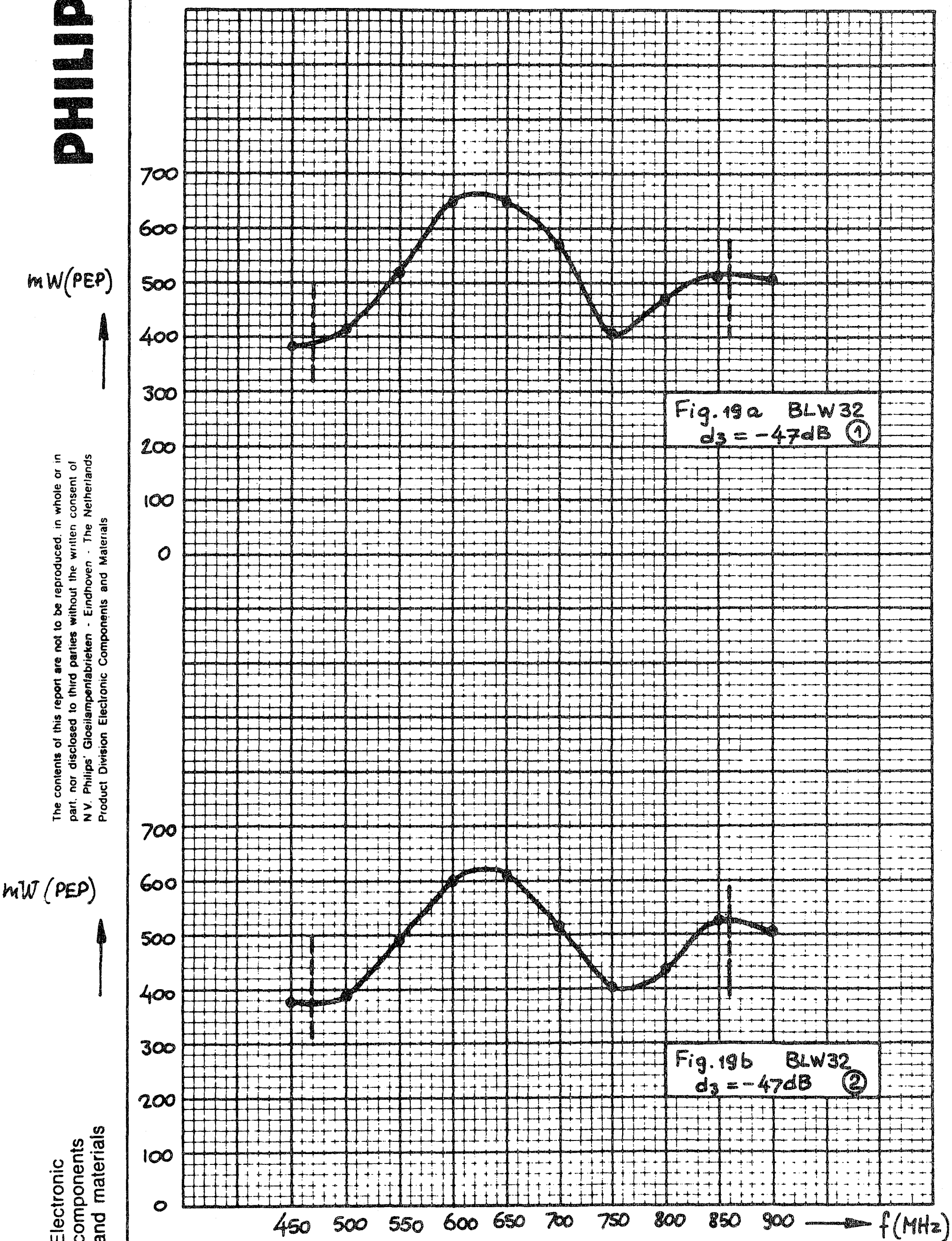

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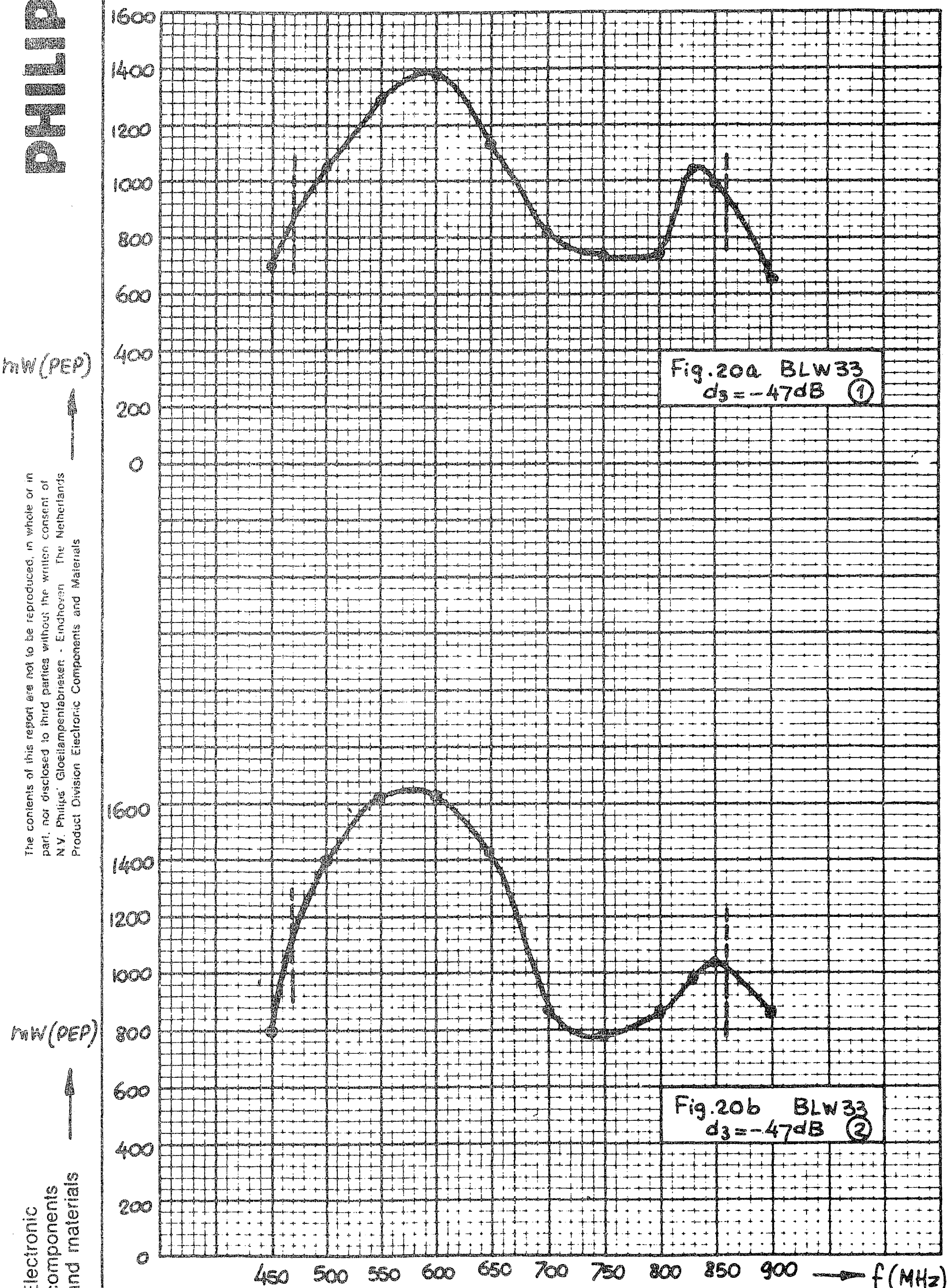

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Electronic
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and materials

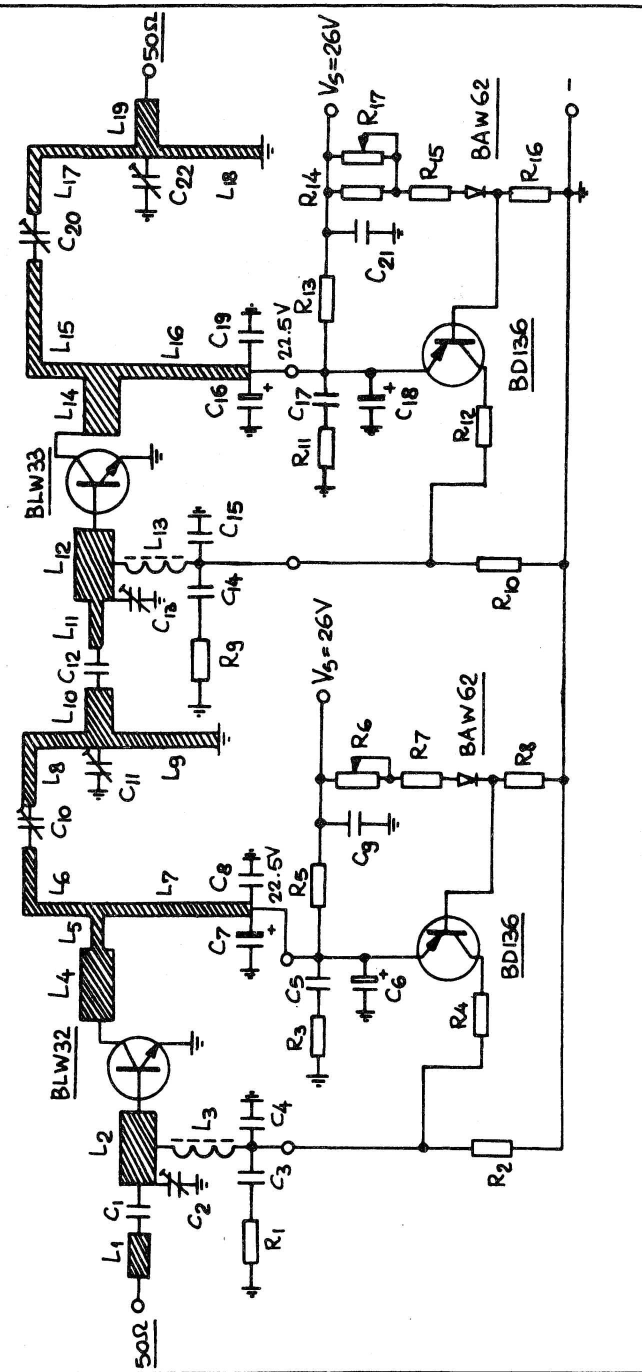

Fig. 22

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8.3. BLW32-BLW33 amplifiers (Figs. 22, 23, 24)

$C_1 = C_{12}$ = 8,2pF, multilayer ceramic chip capacitor, ATC type 100A-8R2-J-Px-50.

$C_2 = C_{11} = C_{13} = C_{22}$ = 1 to 3,5pF film dielectric trimmer (cat. no. 2222 809 05001).

$C_3 = C_9 = C_{14} = C_{21}$ = 100 nF, polyester capacitor.

$C_4 = C_8 = C_{15} = C_{19}$ = 100pF, multilayer ceramic chip capacitor, (cat. no. 2222 852 13101).

$C_5 = C_{17}$ = 470nF, polyester capacitor.

$C_6 = C_7 = C_{16} = C_{18}$ = 6,8 μF, 63V, electrolytic capacitor.

$C_{10} = C_{20}$ = 1 to 4pF, micro thin-trim, Tekelec Airtronic part no. AT 9401-4-SL1.

$L_1 = L_{10} = L_{19}$ = stripline (Z_c = 50Ω), width 4,0 mm.

L_2 = stripline (Z_c = 37,6Ω), 11,5 x 6,0 mm^2.

$L_3 = L_{13}$ = 470nH, microchoke.

$L_4 = L_{14}$ = stripline (Z_c = 37,6Ω), 3,0 x 6,0 mm^2.

L_5 = stripline (Z_c = 102Ω), 4,8 x 1,0 mm^2.

$L_6 = L_8$ = stripline (Z_c = 102Ω), 20,8 x 1,0 mm^2.

L_7 = stripline (Z_c = 102Ω), 19,3 x 1,0 mm^2.

L_9 = stripline (Z_c = 102Ω), 23,4 x 1,0 mm^2.

L_{11} = stripline (Z_c = 102Ω), 9,6 x 1,0 mm^2.

L_{12} = stripline (Z_c = 37,6Ω), 5,3 x 6,0 mm^2.

$L_{15} = L_{17}$ = stripline (Z_c = 102Ω), 21,7 x 1,0 mm^2.

L_{16} = stripline (Z_c = 102Ω), 16,2 x 1,0 mm^2.

L_{18} = stripline (Z_c = 102Ω), 20,4 x 1,0 mm^2.

$R_1 = R_3 = R_9 = R_{11}$ = 10Ω (± 5%) carbon resistor CR25 type.

$R_2 = R_{10}$ = 33Ω (± 5%) carbon resistor CR25 type.

$R_4 = R_{12}$ = 220Ω (± 5%) power metal film resistor PR37 type.

R_5 = 18Ω (± 5%) power metal film resistor PR52 type.

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$R_6 = R_{17} = 220\Omega$, cermet preset potentiometer.
$R_7 = R_{15} = 150\Omega$ ($\pm$ 5%) carbon resistor CR25 style.
$R_8 = R_{16} = 1{,}8k\Omega$ ($\pm$ 5%) carbon resistor CR25 style.
$R_{13} = 10\Omega$ ($\pm$ 5%) enamelled wire-wound resistor WR 0617E style.
$R_{14} = 1k\Omega$ ($\pm$ 10%) carbon resistor CR25 type.

$D_1 = D_2 =$ BAW62.

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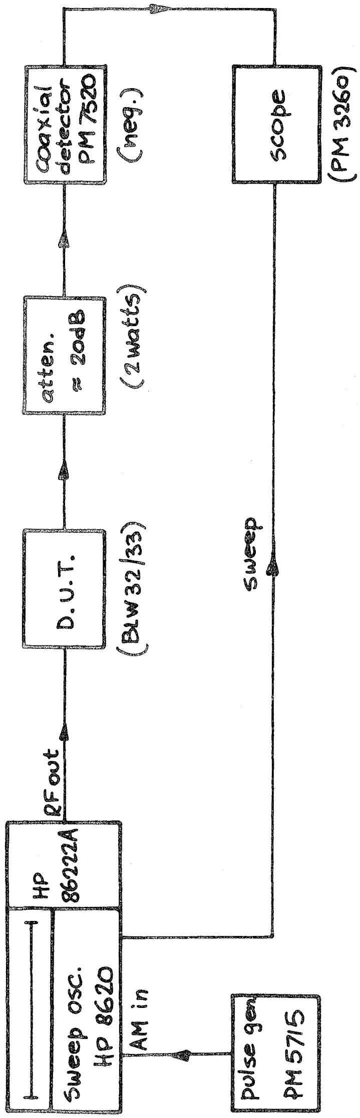
DYNAMIC COMPRESSION TEST SET-UP



rep. time : 20–100 nsec
delay time : 100 nsec
duration time : 10 μsec
ramp time : 6 nsec
output : ± 1.5 V
impedance : $Z_0 = 50\Omega$

Fig. 25

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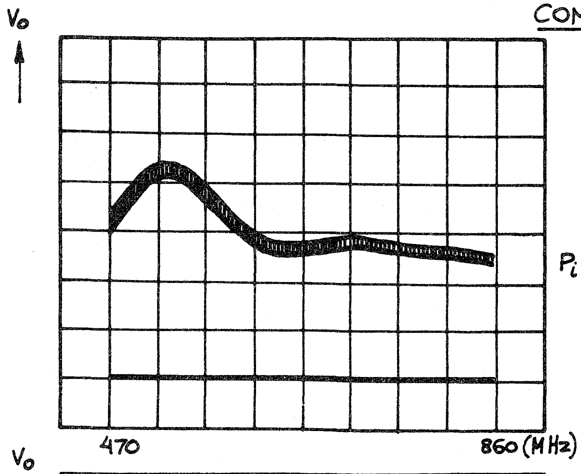
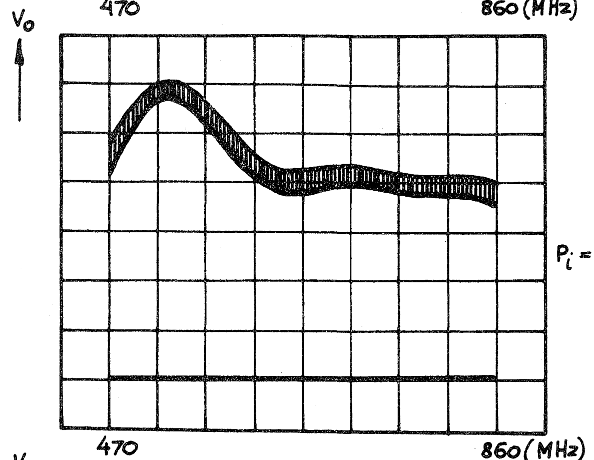
COMPRESSION



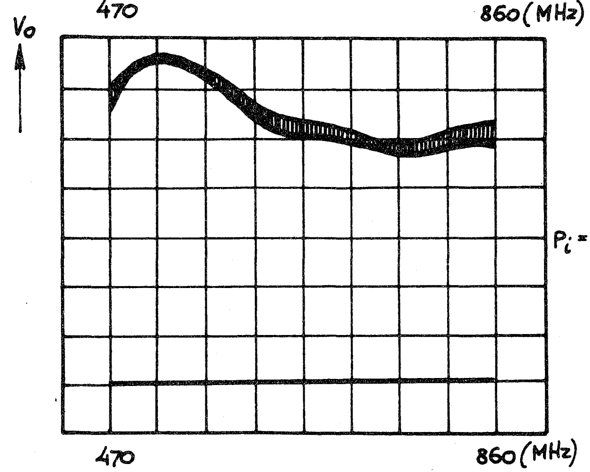
P_i = +7dBm

Fig. 26a



P_i = +10dBm

Fig. 26b



P_i = +13 dBm

Fig. 26c

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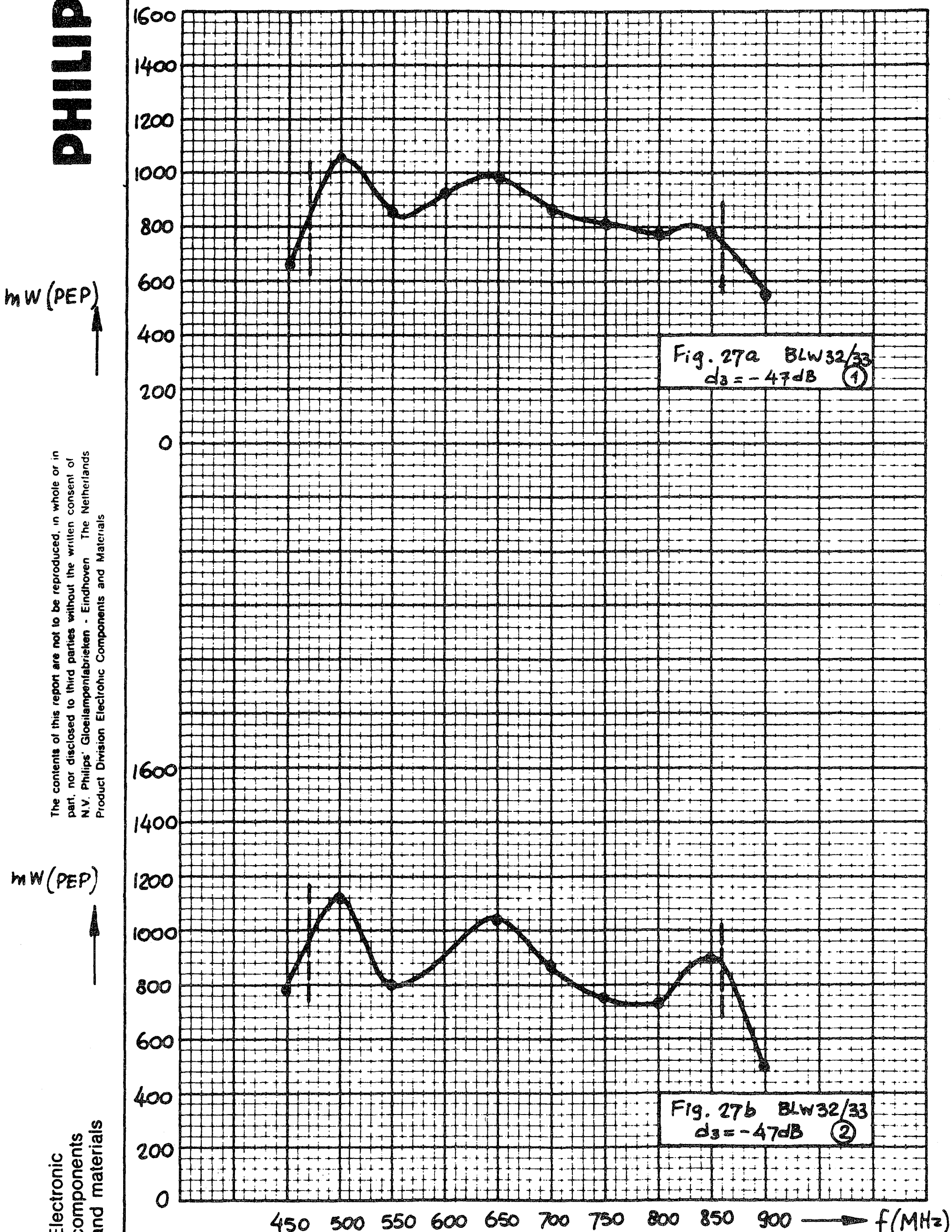

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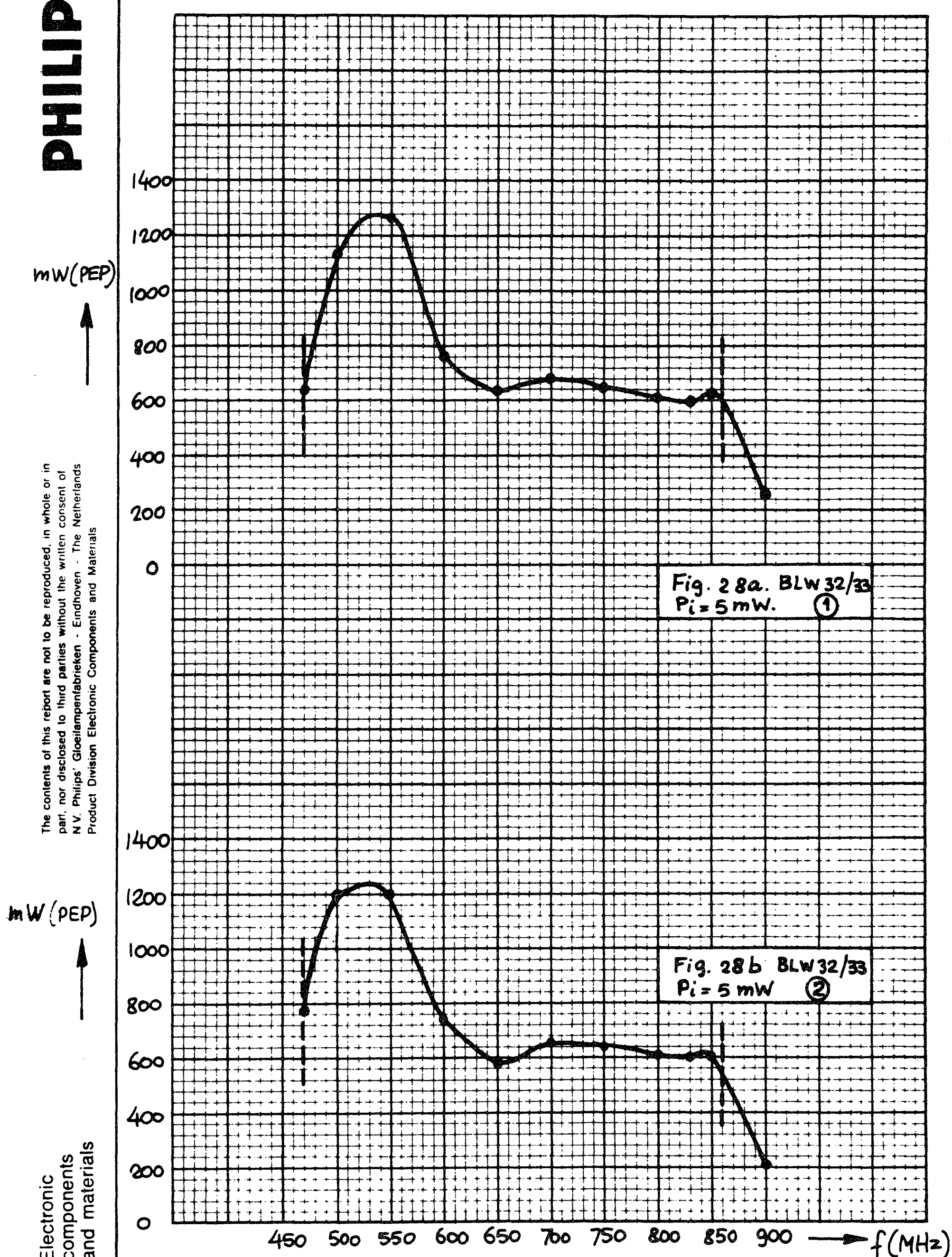

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BLW 32/33

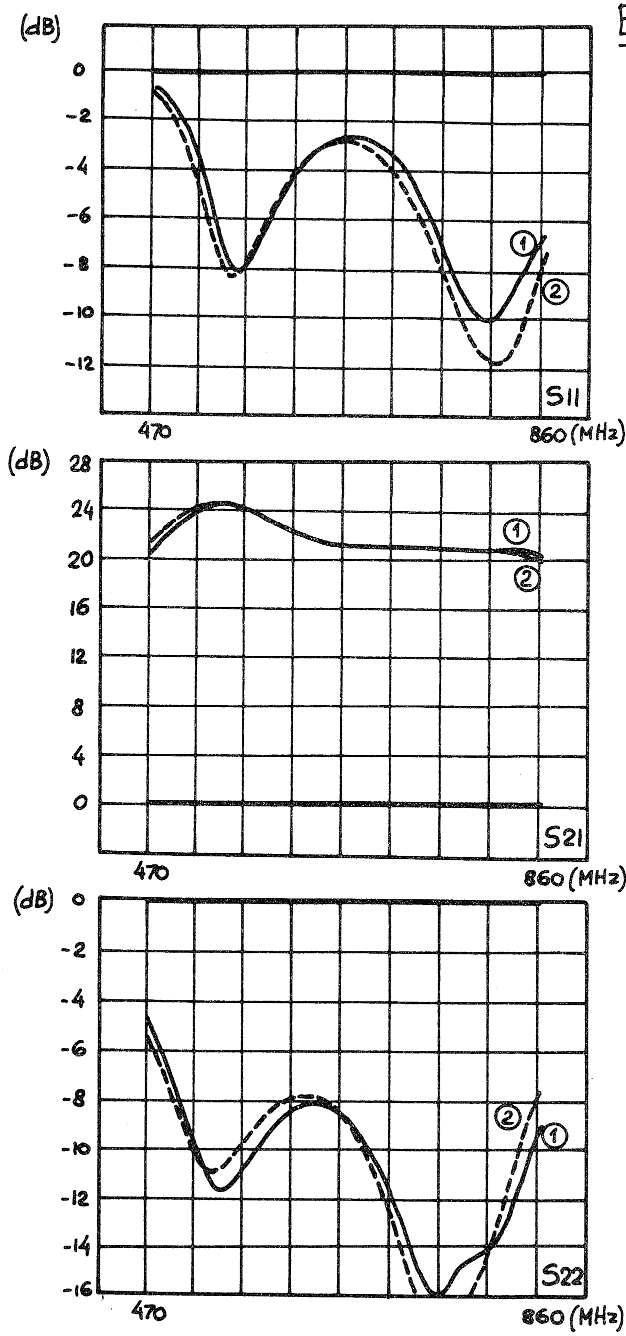

Fig. 29a

Fig. 29b

Fig. 29c

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TABLE 2

BLW32

Channel nr.	i.m.d. (dB)	P_o sync. (mW)		Gain (dB)	
		1*	2	1	2
21	-60	462	427	12,0	12,3
21	-56	605	561		
21	-52	771	716		
39	-60	699	643	11,7	11,15
39	-56	965	937		
39	-52	1245	1189		
70	-60	487	512	10,8	11,3
70	-56	713	769		
70	-52	1007	1021		

* corresponding amplifier number

TABLE 3

BLW33

Channel nr.	i.m.d. (dB)	P_o sync. (mW)		Gain (dB)	
		1	2	1	2
21	-60	1132	1440	10,7	10,3
21	-56	1775	1916		
21	-52	2263	2246		
39	-60	1161	1454	10,6	10,25
39	-56	1762	1902		
39	-52	2293	2321		
70	-60	819	1063	9,5	9,3
70	-56	1342	1580		
70	-52	2014	2265		

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laboratory report

number : ECO 7901 **date** : 20.02.1979

title : A WIDE-BAND LINEAR POWER AMPLIFIER (470-860 MHz) WITH TWO TRANSISTORS BLW34

author : A.H. Hilbers and M.J. Köppen

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laboratory report

central application laboratory CAB
eindhoven - the netherlands

number: ECO 7901 date: 20.02.1979

project: 6823 pages: A1 ; S6 ; R25

title A WIDE-BAND LINEAR POWER AMPLIFIER (470-860 MHz) WITH TWO TRANSISTORS BLW34

authors A.H. Hilbers and M.J. Köppen

ABSTRACT:

For application in driver or final stages of TV-transposers in band IV/V (470-860 MHz) a linear wideband power amplifier has been designed with 2 transistors BLW34 coupled by means of 3dB-90^{o} hybrids.

Each transistor is adjusted in class-A at V_{CE} = 25V and I_c = 0,6A.

The peak sync output power for a 3-tone I.M. distortion of -60dB varies between 3,6 and 5,4W.

The power gain is 9,1 $\pm$ 0,3dB.

Input and output VSWR are below 1,3.

Appr. R.A. Pölzl

Advies Octrooi d.d. 13 mrt. 1979	X AV	GV		B		BL	
Opgave Mamo d.d. - 2 mrt. 1979	X AV	GV	X SP	B		BL	
Datum: 26 feb. 1979	Mamo						

SUMMARY

Fig. A shows the basic circuit of the prototype with two transistors BLW34 both operating in class A.

The amplifier consists of two equal branches. Both are coupled at the input and output side by means of 3dB-90^{0} coaxial hybrids (500-1000MHz).

The circuit is mounted on a double-clad printed circuit board. To keep the losses in the transforming strip transmission lines sufficiently low the dielectric is PTFE glass fibre with an ε_r = 2,74 having a thickness of 1/32 inch. The fixed capacitors of the r.f. chain are of the ceramic multilayer chip type.

To obtain a high degree of linearity (-60dB i.m.d., three specified tones) the transistors have to operate in class A. This class A adjustment is controlled by means of the circuit of Fig. B. Each branch needs such a supply unit.

The circuit contains all typical elements to assure stable and spurious-free operation even under mismatched input and output conditions. The applied decoupling elements have been chosen so that the effect on the video step response is very small.

The theoretical design started with the output side in which the optimum load impedance (class A) varied between 12,8 + j11,0 ohms at 470MHz and 5,4 + j 7,7 ohms at 860MHz.
The output capacitance of the BLW34 is tuned out with the collector choke in the middle of the band.
Further, wide-band transformation to the 50 ohms amplifier load is arranged with the aid of a Chebyshev transforming network.

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On the transistor input the impedance ranges from 1,5 + j1,9 ohms at 470 MHz to 1,3 + j4 ohms at 860 MHz, whilst the power gain varies with appr. 5dB/octave.

Reducing the gain variation based on permitting input mismatch at the lower frequencies has been done according to a method described in Ref. 1.

The design has been theoretically optimized with the aid of a computer.
For practical optimization and tuning the amplifier was inserted in a dynamic gain compression test set-up.

Fig. C shows the small signal input and output reflection damping (S_{11} and S_{22} , whilst the power gain versus frequency is represented by the S_{21} curve.
These results are given for two heatsink temperatures viz. 23 and 70°C.

The output power $P_{o\ sync}$ versus frequency for a three-tone linearity of -60dB is shown in Fig. D.

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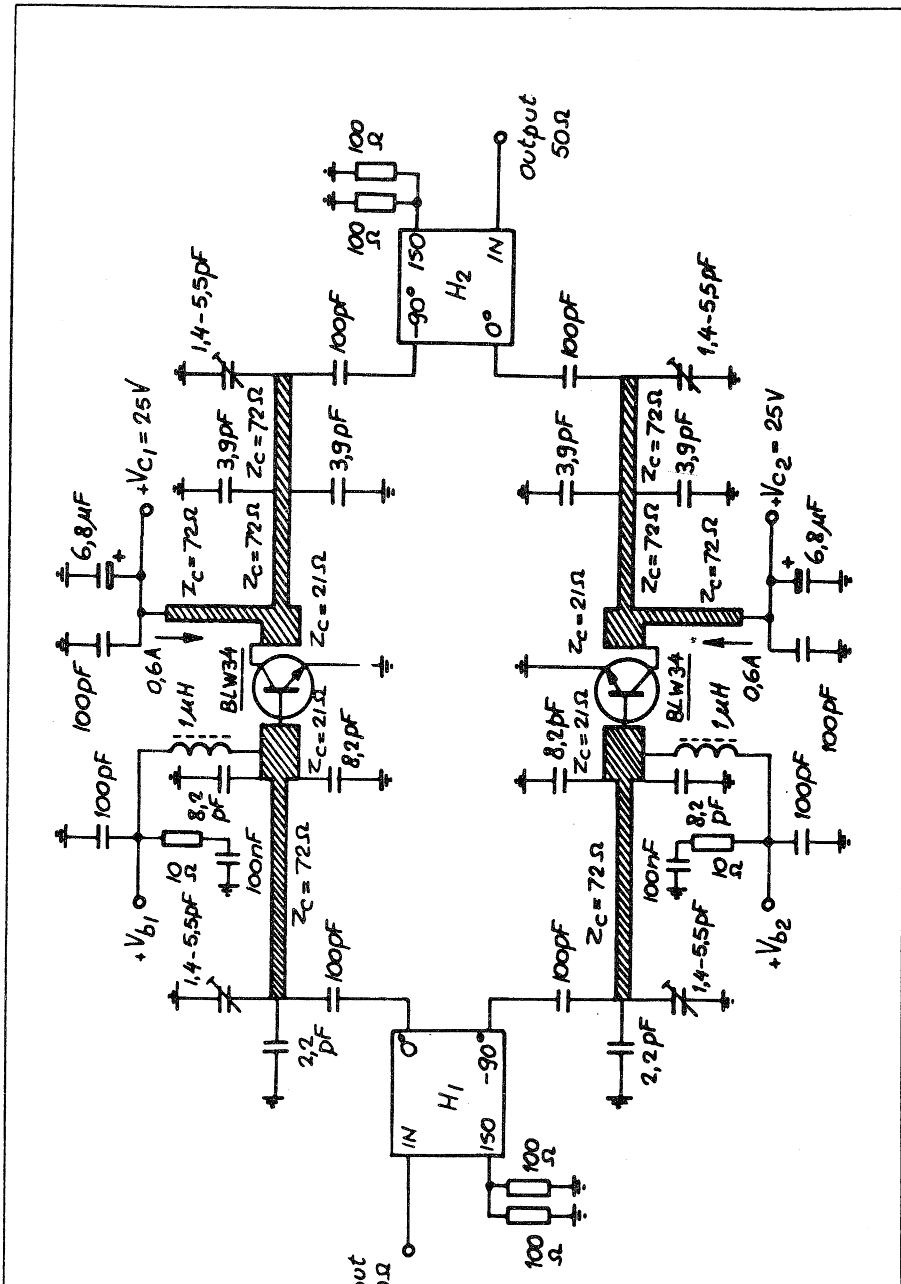

Fig. A

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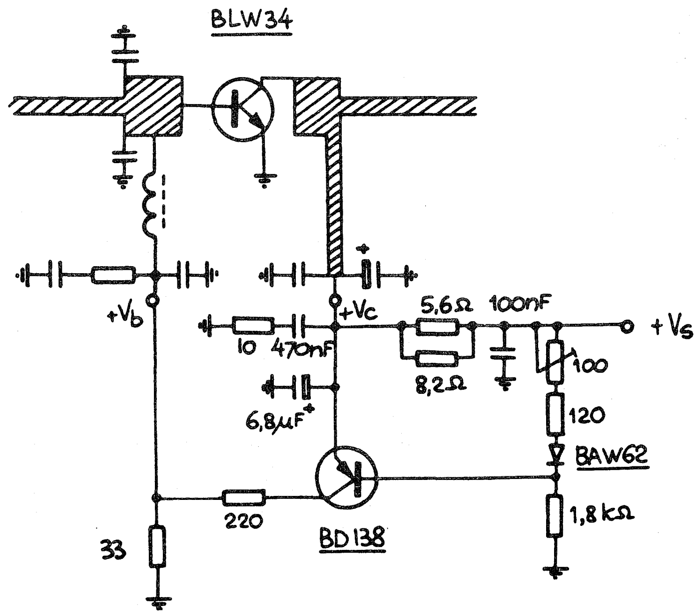

Fig. B

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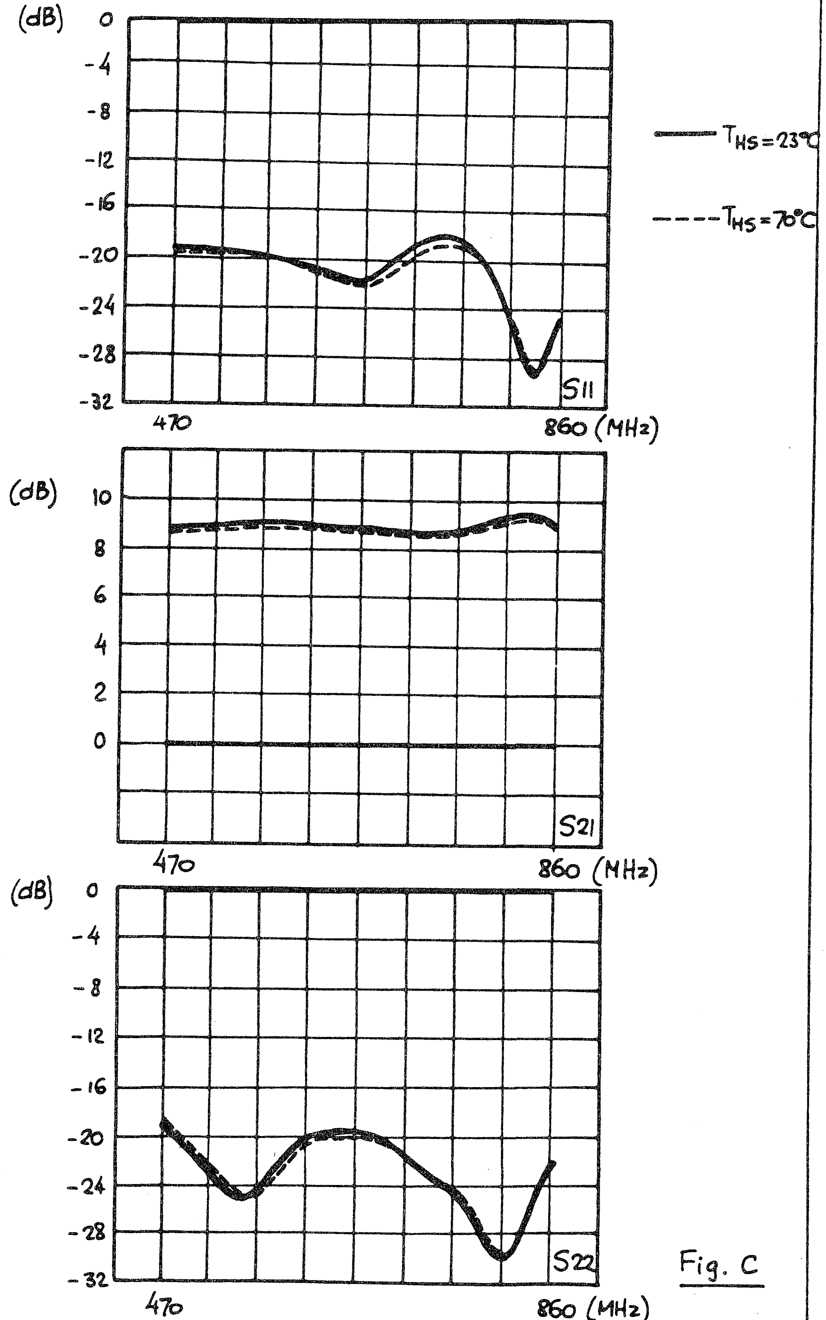

Fig. C

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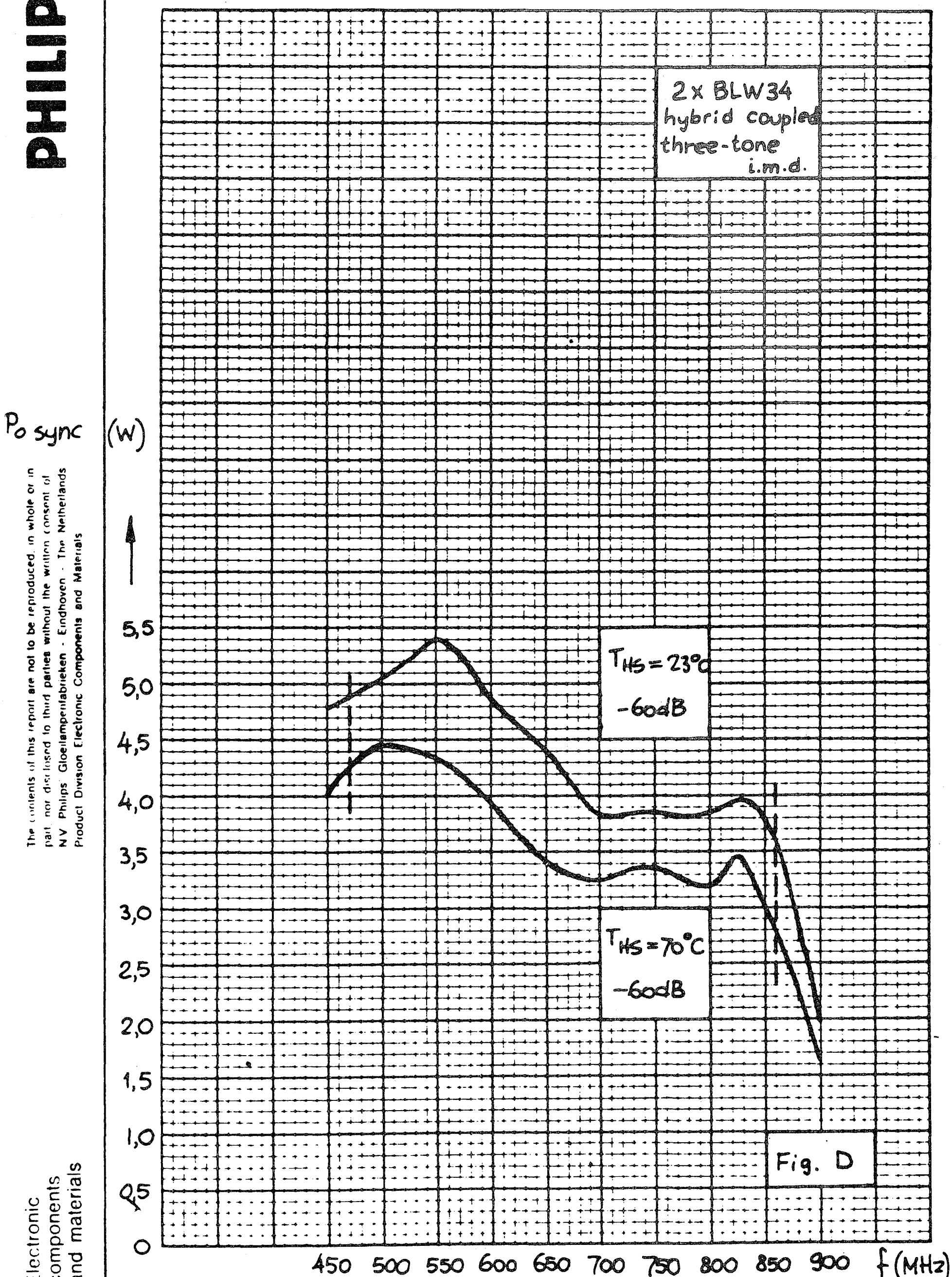

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1. INTRODUCTION

This report describes the realisation of a wide-band UHF power amplifier for TV transposer service in band IV and V (470-860MHz).

The amplifier is designed with the new BLW34 transistors being developed for ultra linear applications operating in class A.

Each device is able to deliver at least 1.8W peak sync·output. The BLW34 forms a series with the smaller devices BLW 33 (1,0W) and BLW32 (0,5W).

The power gain at 860MHz is at least 9dB.

The BLW34 is encapsulated in a ¼ inch capstan envelope with ceramic cap.

2. THEORETICAL CONSIDERATIONS

2.1. The equivalent circuit of the BLW34

For class A operation the BLW34 is specified at V_{CE} = 25V; I_C = 600mA.

The corresponding typical gain, input and load impedances according to the Data sheets are given below:

f	gain	R_i (series)	X_i (series)	R_L (series)	X_L (series)
(MHz)	(dB)	(Ohm)	(Ohm)	(Ohm)	(Ohm)
470	15,2	1,46	1,93	12,8	11,0
636	12,6	1,39	2,84	8,85	9,97
860	10,1	1,27	4,00	5,36	7,67

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To facilitate calculations an approximate equivalent circuit for the transistor input and output impedance can be given. It is shown in Fig. 1.

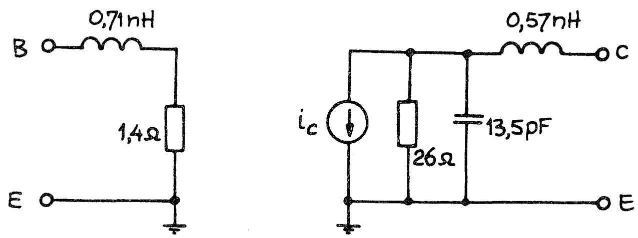

Fig. 1.

2.2. The output network

The circuit will be designed on printed circuit board with PTFE fibre glass as a dielectric having an ε_r = 2,74 and thickness of 1/32 inch.

The input and output network start with a piece of stripline having a width of 6mm, being the width of the base and collector leads. For a dielectric of 1/32 inch the characteristic resistance is 21 ohms.

The length for the collector lead amounts to 3 mm, but the base lead is different in length.

The first step in the matching is to tune out the output capacitance of the transistor by means of the collector R.F. choke.

This choke is executed as a stripline with a width of 1 mm corresponding with a characteristic resistance of 72 ohms to keep the parallel capacitance at this point as low as possible.

For practical reasons the choke is connected to the main transmission line at a distance of 3mm from the transistor edge. The design procedure is practically the same as described in report no. ECO 7704 (Ref. 2) except for some small differences:

a. the length of the collector terminal has been reduced to 3mm to decrease parasitic capacitance at this point to a minimum.

b. the characteristic resistance of the series lines has been increased to 72 ohms to reduce the size of the p.c. board.

The results of the calculations before and after computer optimization are given in Fig. 2 and table I.

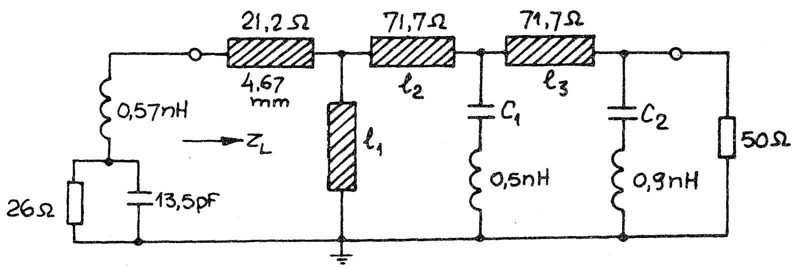

Fig. 2.

element	before optim.	after optim.	unit
l_1	16,8	22,1	mm
l_2	18,2	20,8	mm
C_1	8,58	9,76	pF
l_3	38,2	43,5	mm
C_2	3,91	3,09	pF
$S_{max.}$	2,48	1,23	-

Table I.

S_{max} is the maximum VSWR of the network.

The lenghts given hold for air lines. The actual lengths on the p.c. board are shorter; the reduction factors are; 1,445 for the 72 ohms lines and 1,556 for the 21 ohms line.

The predicted minimum output power of the complete amplifier with two transistors BLW34 is:

$$P_{o_2} = \frac{2P_{o_1}}{S_{max}} \cdot 0{,}95 = \frac{2 \cdot 1{,}8}{1{,}23} \cdot 0{,}95 \approx 2{,}8 \text{ Watts}$$

P_{o_1} is the minimum output power of one transistor in a narrow band circuit, S_{max} is as specified above and the factor 0,95 represents the power loss of the hybrid coupler at the output.

2.3. The input network

The design procedure is again almost equal to that given in Ref. 2. The only small differences are:

a. the length of the base terminal has been reduced such that it has just the correct inductance for the first matching section. The width is 6mm, corresponding with an Rc of 21 ohms.

b. the Rc of the other series line has been increased to 72 ohms for the same reason as in the output network.

The results of the calculations are summarized in Fig. 3 and table II.

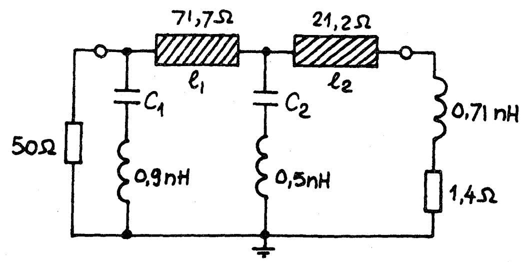

Fig. 3.

element	before optim.	after optim.	unit
C_1	4,78	4,41	pF
l_1	25,9	32,1	mm
C_2	21,9	19,9	pF
l_2	13,3	10,5	mm
ΔG	$\pm 1,93$	$\pm 0,12$	dB

Table II

ΔG is the resulting power gain variation caused by the transistor and the input network over the frequency band. The lengths of the lines hold for air as a dielectric. Transformation to striplines on a p.c. board is done in the same way as in the previous section.

The minimum power gain of the complete amplifier with two transistors BLW34 is expected to be:

$$G_o = G_t - 2A_H - A_1 - A_2$$

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in which: G_t = minimum power gain of BLW34 in a narrow band circuit.

A_H = power loss of one hybrid coupler

A_1 = reflection loss of input network

A_2 = reflection loss of output network

In practical figures this means:

$$G_0 = 9{,}0 - 2.\ 0{,}2 - 0{,}15 - 0{,}05 = 8{,}4\text{dB}$$

The input VSWR of a single amplifier was calculated to vary from 11 at 470 MHz down to 1,45 at 860 MHz.

3. THE HYBRID COUPLED AMPLIFIER

3.1. Practical considerations

On previous pages the theoretical approach of a single amplifier has been discussed.

In practice, it was the intention to realize a small compact amplifier on a p.c. board with the input and output terminal (Rc = 50 ohms) in line for easy cascading of several amplifiers.

Besides the wide-band properties it is the intention to obtain a higher output power, so two BLW34 branches are connected in parallel.

At the same time it is of course rather unacceptable that the amplifier loads a driver stage with a mismatch causing a VSWR of 11 at 470MHz.

Both problems may be solved sufficiently by applying two BLW34 branches in parallel with the aid of two wide-band 3dB-90^{0} coaxial hybrids on a 50 ohms basis.

In this configuration the output power will be nearly doubled; reduction of the input VSWR of the complete system to a value of around 1,2 may be explained from the properties of the coupler applied. The reflected power will be absorbed in the resistor matching the isolated port. This resistor is 50 ohms and consists of two 100 ohms power metal film resistors in parallel. The same has been done on the output side.

The p.c. board needs to be double copper clad with a PTFE fibre glass dielectric (ε_r = 2,74) for low losses at UHF. A tnickness of 1/32 inch has been chosen; so the 72 ohms lines are 1mm wide.

Fig. 4 shows the circuit diagram of the complete 2xBLW34 class A amplifier. The biasing circuit is drawn in Fig. 5.

The p.c. board is in Fig. 6 and the amplifier lay-out in Fig. 7.

For a correct earthing the upper earth sheet parts are connected to the lower sheet by soldering copper straps at the edges of the p.c. board. The black parts in Fig. 7 are the soldered copper straps.

The emitters are grounded as short as possible by applying copper straps under the emitter leads. For that reason the holes in the board are square instead of round.

Both transistors are screwed to a water-cooled brass heatsink. So, several heatsink temperatures can be applied by means of a TAMSON unit supplying water with a thermostatically controlled temperature.

The tuning capacitors in the circuit are of the film dielectric type with three tags of which both earth terminals are fed through small holes and soldered to the upper as well as the lower plane. Fixed capacitors in the r.f. path are of the multi-layer ceramic chip type.

The coaxial connectors are of the SMA 50 ohms type, being soldered to upper and lower sheet.

3.2. Practical optimization

We started with optimization on a small signal basis with the circuit inserted in a network analyzer chain, having swept S-parameter facilities.

So far, similar tuning methods have been applied as described in Refs. 2 and 4.

Because it is rather complicated to find the best compromise between an acceptable flat gain curve (S_{21}) and sufficient output power with low i.m.d. a dynamic large signal optimization method has been realized. Fig. 8 shows the block diagram.

This tuning method is based on the correlation between the single tone 1dB compression point and the i.m.d. figure of a linear amplifier.

In this set-up the swept output power level of the amplifier under test is kept constant and the required (detected) drive power monitored on an oscilloscope screen (PM3260).

The swept drive power is available from the sweep generator HP8620C in combination with RF plug-in unit HP86222A. Because the output power of this system is too low viz. appr. 20mW (+13dBm) a combined amplifier with BLW32 and BLW33

(Ref. 4) has been added. The latter combination shows an overall gain of approx. 20dB in the range 470-860MHz.
When the circuit under test is inserted in the chain of Fig. 8, the input power measured on port C of hybrid 1 corresponds in principle with the gain curve (S_{21}) being measured with the network analyzer; in fact one is the inverse of the other.

When the drive level is slowly increased, the shape of the gain curve changes somewhat when compression starts.
By careful retuning of the amplifier the shape of the gain curve can be corrected again in the direction of the original smaller signal curve.

The actual single tone output power has been measured with the aid of the calorimetric watt meter HP435A when the action of the sweep is stopped. Also it is possible to examine the output signal itself by means of a spectrum analyzer being loosely coupled via a 50 ohms pick-off device.

Resuming it can be said that the advantage of applying this high level tuning system is characterized by the fact that the output power is leveled and so compression does not start earlier due to gain fluctuations of the amplifier.
This makes the judgement of the compression level easier.

3.3. Intermodulation, VSWR and gain measurements

For i.m.d. measurements on television systems the post offices advise and apply the 3 tone test method (vision carrier -8dB, sound carrier -7dB, sideband signal -16dB; zero dB corresponds to peak sync level).

For this reason a wideband test set-up has been realized.
The block diagram is in Fig. 9.

In this set-up first the sound and vision carriers are joined in the wide-band coaxial hybrid H_1.
Then the sideband signal from the (smaller) generator SMLU is added in hybrid H_2.

The complete three-tone signal passes a low-pass filter (700 or 1000 MHz cut-off depending on input frequency) a continuously variable attenuator and a circulator (three different types needed to cover at least the range 470-860 MHz).

The output power is measured with a HP435A calorimetric watt meter and the i.m.d. observed with a spectrum analyzer (HP8558B).

Fig. 10 shows the three-tone i.m.d. results measured on the complete hybrid coupled 2xBLW34 amplifier. To get an idea of the $P_{o\ sync}$ drop for different heatsink temperatures, measurements have been done for $T_{HS} = 23^{o}C$ and $T_{HS} = 70^{o}C$. The maximum drop in $P_{o\ sync}$ amounts to 1,1dB.

Besides the three-tone test the peak envelope power has been measured in a two-tone way for a third order i.m.d. of $d_3 = -47dB$. The two-tone test set-up corresponds to the one described in Ref. 3.

It can be proved that there is a correlation between the afore described -60dB three-tone and the -47dB two-tone test method.

It is known that the advantage of video precorrection on i.m.d. amounts to values up to 10dB. Calculating with an average of 8dB, the two-tone i.m.d. for d_3 = 39dB is interesting.
From Fig. 11 it may be seen that the output power is more than doubled. Both tests have been done for heatsink temperatures of resp. 23 and 70^{o}C.

Finally input and output VSWR and gain figures were measured only under small signal conditions. The results of amplifier branches 1 and 2 for a heatsink temperature of 70^{o}C are shown in Figs. 12a, b, c. The VSWR figures of the input and output are expressed in reflection damping (resp. S_{11} and S_{22}) on a 50 ohms basis.

According to Fig. 12a the minimum reflection damping of the input amounts to S_{11} = -1dB (VSWR = 17,4).at 470MHz and to S_{11} = -10dB (VSWR = 1,93) at 860MHz.

On the output side the worst reflection damping amounts to S_{22} =-5,5dB (VSWR = 3,26) in the middle of the passband (Fig. 12c).

The power gain, represented by S_{21}, is appr. 9,5dB (Fig. 12b).

Resuming it can be said that, as Fig. 12 shows, the differences in gain, input and output VSWR between both branches are rather small.

Figs. 13a, b, c show the final results of the complete hybrid coupled amplifier. Examining the S_{11} and S_{22} figures it appears that the minimum input reflection damping (S_{11}) amounts to appr. 18dB, what corresponds with a maximum VSWR of 1,29.

The latter value may be mainly explained from the specified maximum VSWR = 1,25 of the applied Anaren hybrids.
If more attention is paid to matching of the isolated ports the result may be improved. The matching consists of two metal film power transistors of 100 ohms in parallel.
They have low-frequency tolerances of 5%.

The small signal gain of the complete amplifier amounts to 9,1 $\pm$ 0,3dB for $T_{HS} = 23^{o}C$, whilst the dashed line shows the results for $T_{HS} = 70^{o}C$ being 8,9 $\pm$ 0,3dB.

The temperature influence on the S_{11} and S_{22} figures is almost negligible.

4. CONCLUSIONS

On preceding pages the theoretical and practical design has been described of a wide-band (470-860 MHz) high quality linear amplifier being equipped with two BLW34 transistors operating in class A.

There are some small differences between the theoretical design and the practical circuit.

- The calculated value for the chip capacitors $C_{11} = C_{12} = C_{13} = C_{14}$ in Fig. 4 was 10pF. In practice 8,2pF appeared to be a better choice.

- Also the values of $C_{19} = C_{20} = C_{21} = C_{22}$, being calculated as 4,7pF are changed. The new value is 3,9pF.

- In the first instance a chip capacitor of 1,5pF in parallel with C_{23} and C_{26} was planned. However this capacitor could be omitted in the operational circuit.

The expected results for a single amplifier: $G_{p\ min.} = 8{,}8dB$; and $P_{o\ sync} = 1{,}5W$ at a three-tone i.m.d. of -60dB and $G_{p\ min} = 8{,}4dB$ and $P_{o\ sync.} = 2{,}8W$ for $T_{HS} \leq 70^{o}C$ are realized in this design.

Here, we have calculated with a total insertion loss of 0,4dB for the applied hybrids. They are specified for a maximum of 0,25dB per device (1,06 x power).

5. REFERENCES

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IEEE Transactions on microwave theory and techniques, Vol. MTT-19, no. 4, April 1971, pp. 381-386.

Ref. 2: A.H. Hilbers and M.J. Köppen - Design of a wide band linear power amplifier (470-860MHz) with two transistors BLW98. C.A.B. report ECO 7704.

Ref. 3: M.J. Köppen - The BLX98 as a linear amplifier at 1GHz. C.A.B. report ECO 7601.

Ref. 4: A.H. Hilbers and M.J. Köppen - Wide-band linear power amplifiers (470-860MHz) with the transistors BLW32 and BLW33. C.A.B. report ECO 7806.

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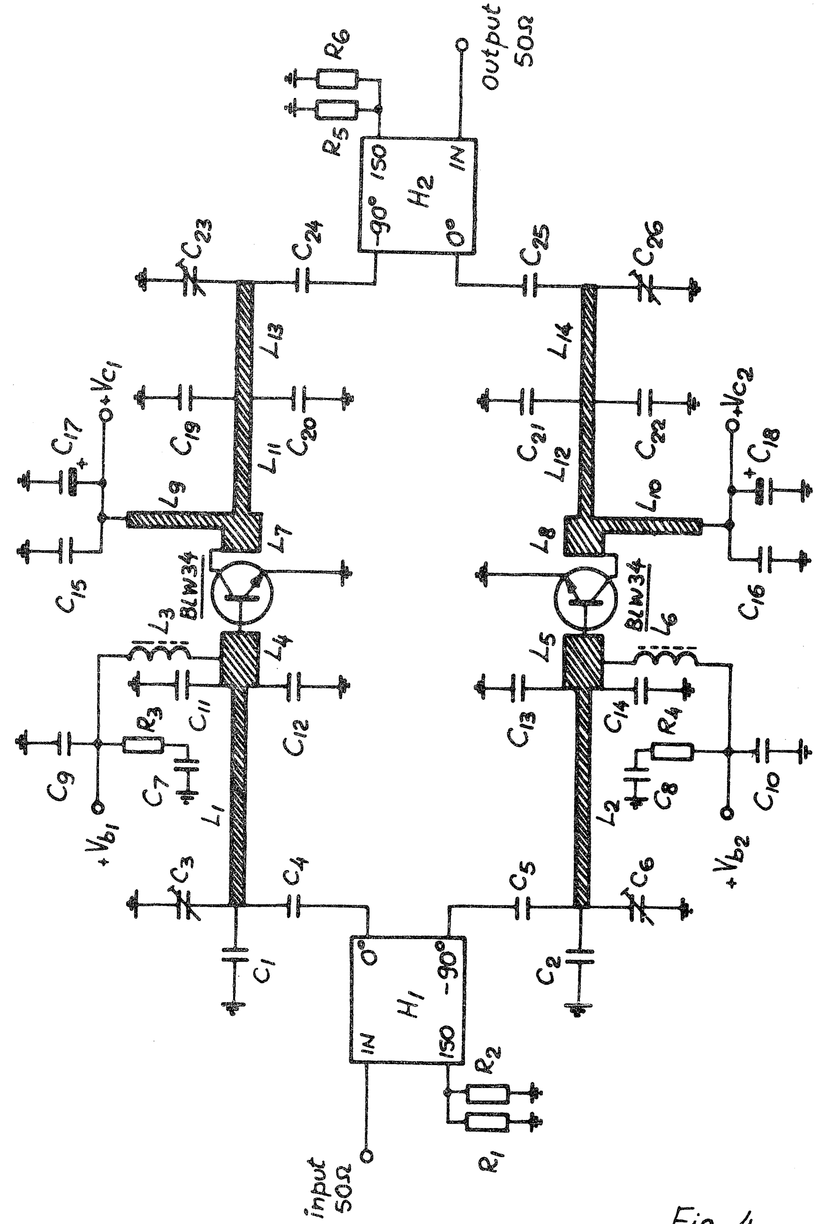

Fig. 4

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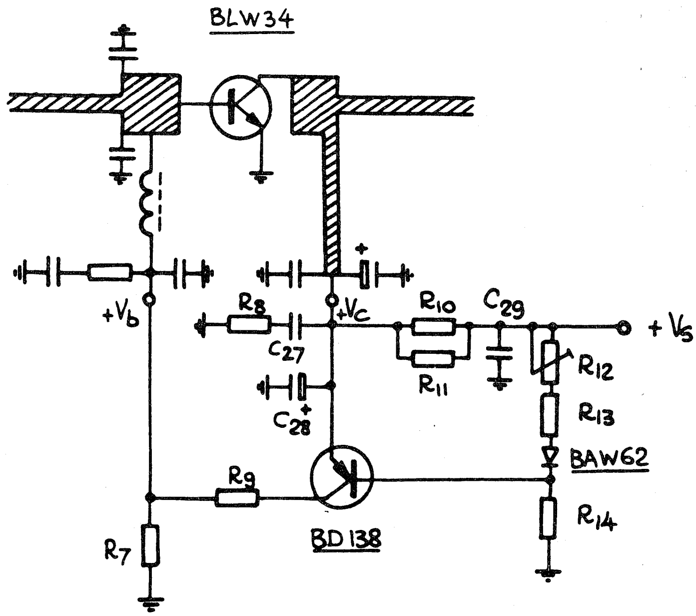

Fig. 5

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6. LIST OF COMPONENTS

$C_1 = C_2$ = 2,2pF, multilayer ceramic chip capacitor.
Tekelec-Airtronic part no. 500 R15 N2R2 BA

$C_3 = C_6 = C_{23} = C_{26}$ = 1,4 to 5,5pF film dielectric trimmer (cat. no. 2222 809 09001).

$C_4 = C_5 = C_{24} = C_{25}$ = 100pF, multilayer ceramic chip capacitor (cat. no. 2222 851 13101).

$C_7 = C_8 = C_{29}$ = 100nF, polyester capacitor.

$C_9 = C_{10} = C_{15} = C_{16}$ = 100pF, multilayer ceramic chip capacitor (cat. no. 2222 852 13101).

$C_{11} = C_{12} = C_{13} = C_{14}$ = 8,2pF, multilayer ceramic chip capacitor, ATC (American Technical Ceramics) type 100A-8R2-J-Px-50.

$C_{17} = C_{18} = C_{28}$ = 6,8μF, 63V, electrolytic capacitor.

$C_{19} = C_{20} = C_{21} = C_{22}$ = 3,9pF, multilayer ceramic chip capacitor, Tekelec-Airtronic part no. 500 R15 N3R9 CA.

C_{27} = 470nF, polyester capacitor.

$R_1 = R_2 = R_5 = R_6$ = 100Ω (± 5%), power metal film resistor PR37 type, (cat. no. 2322 191 31001).

$R_3 = R_4 = R_8$ = 10Ω (± 5%) carbon resistor; CR25 type.

R_7 = 33Ω (± 5%), carbon resistor; CR25 type.

R_9 = 220Ω (± 5%), power metal film resistor PR52 type (cat. no. 2322 192 32201).

R_{10} = 5,6Ω (± 5%), enamelled wire-wound resistor WR0617E style.

R_{11} = 8,2Ω (± 5%), enamelled wire-wound resistor WR0617E style.

R_{12} = 100Ω, cermet preset potentiometer.

R_{13} = 120Ω (± 5%), carbon resistor; CR25 type.

R_{14} = 1,8kΩ (± 5%), carbon resistor; CR25 type.

$H_1 = H_2$ = ultra-miniature 3dB-90^o coupler
model no. 10264-3, range 0,5-1,0GHz.
Anaren Microwave Inc.

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$L_1 = L_2$ = stripline (Zc = 72Ω), 22,1 x 1,0mm^2 *
$L_3 = L_6$ = 1μH; microchoke
$L_4 = L_5$ = stripline (Zc = 21Ω), 6,7 x 6,0mm^2 *
$L_7 = L_8$ = stripline (Zc = 21Ω), 3,0 x 6,0mm^2 *
$L_9 = L_{10}$ = stripline (Zc = 72Ω),15,2 x 1,0mm^2 *
$L_{11} = L_{12}$ = stripline (Zc = 72Ω), 14,3 x 1,0mm^2 *
$L_{13} = L_{14}$ = stripline (Zc = 72Ω), 29,9 x 1,0mm^2 *

* These striplines are printed on double Cu-clad printed circuit board with PTFE fibre-glass dielectric (ε_r = 2,74); thickness 1/32".

../GS

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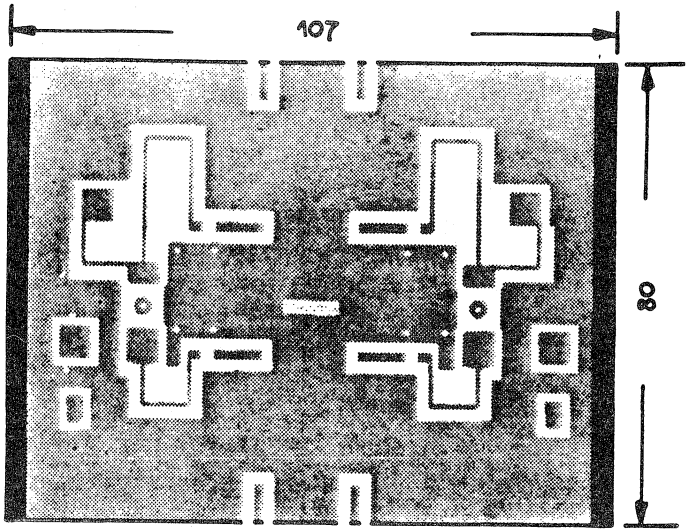
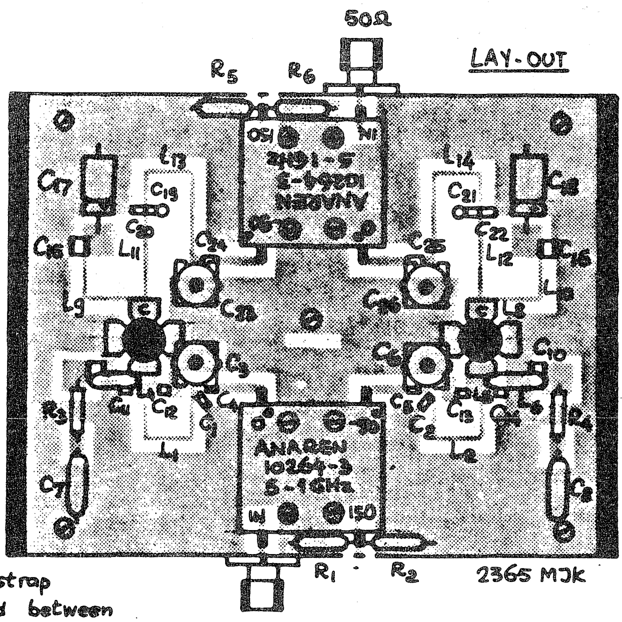


P.C. BOARD
1/32 inch PTFE
double Cu clad

Fig. 6

LAY-OUT



2365 MJK

copper strap
soldered between
upper and lower sheet.

Fig. 7

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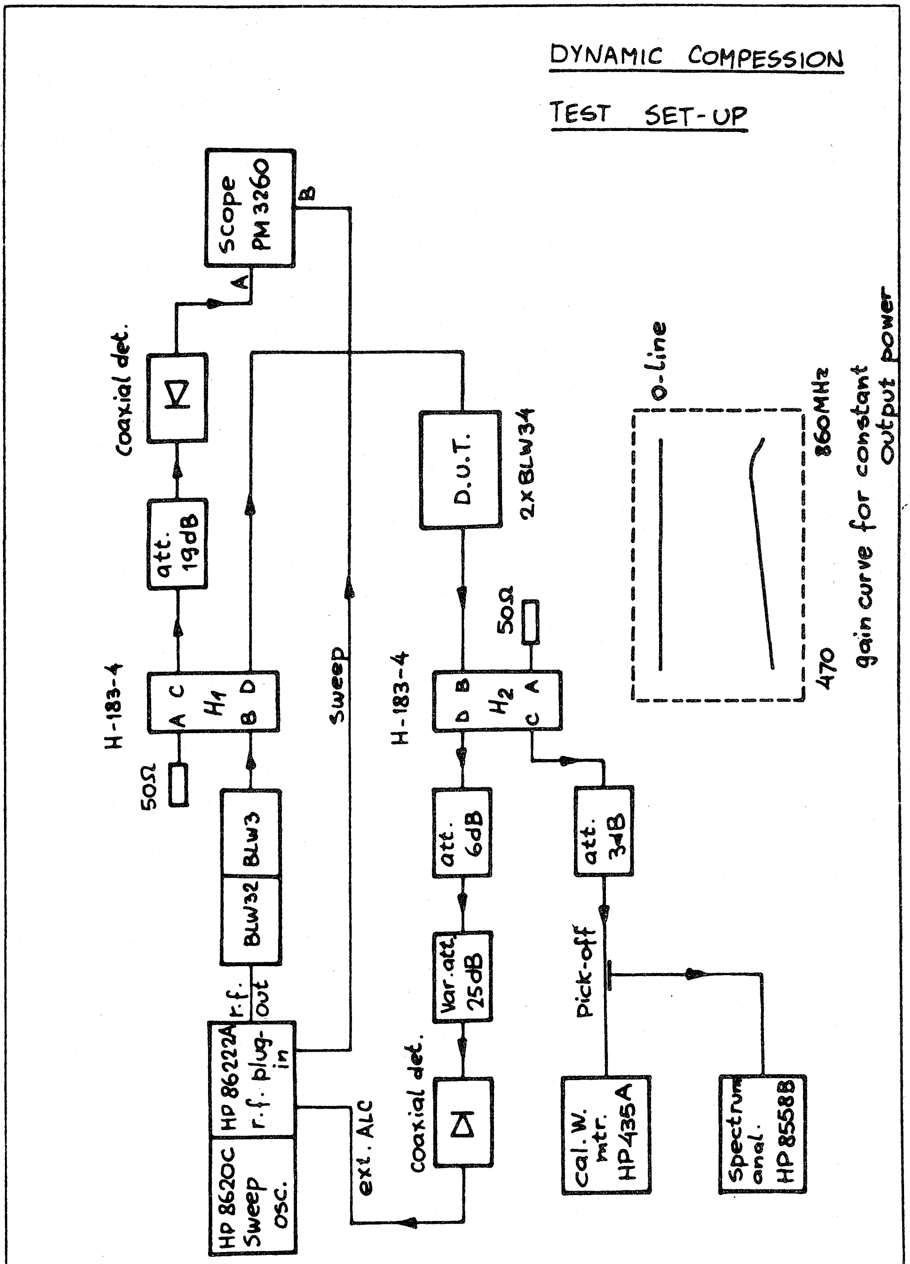

Fig. 8

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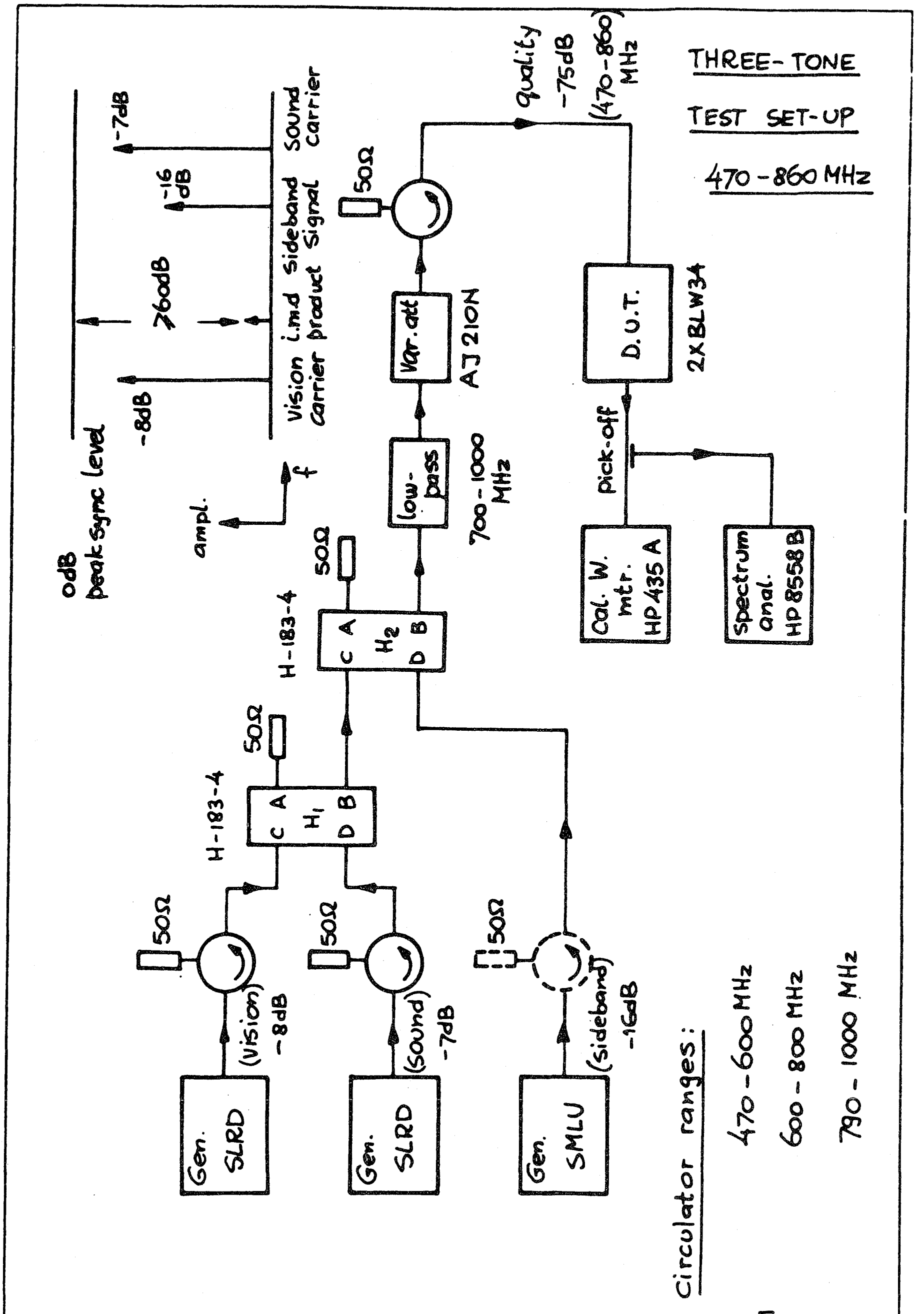

Fig. 9

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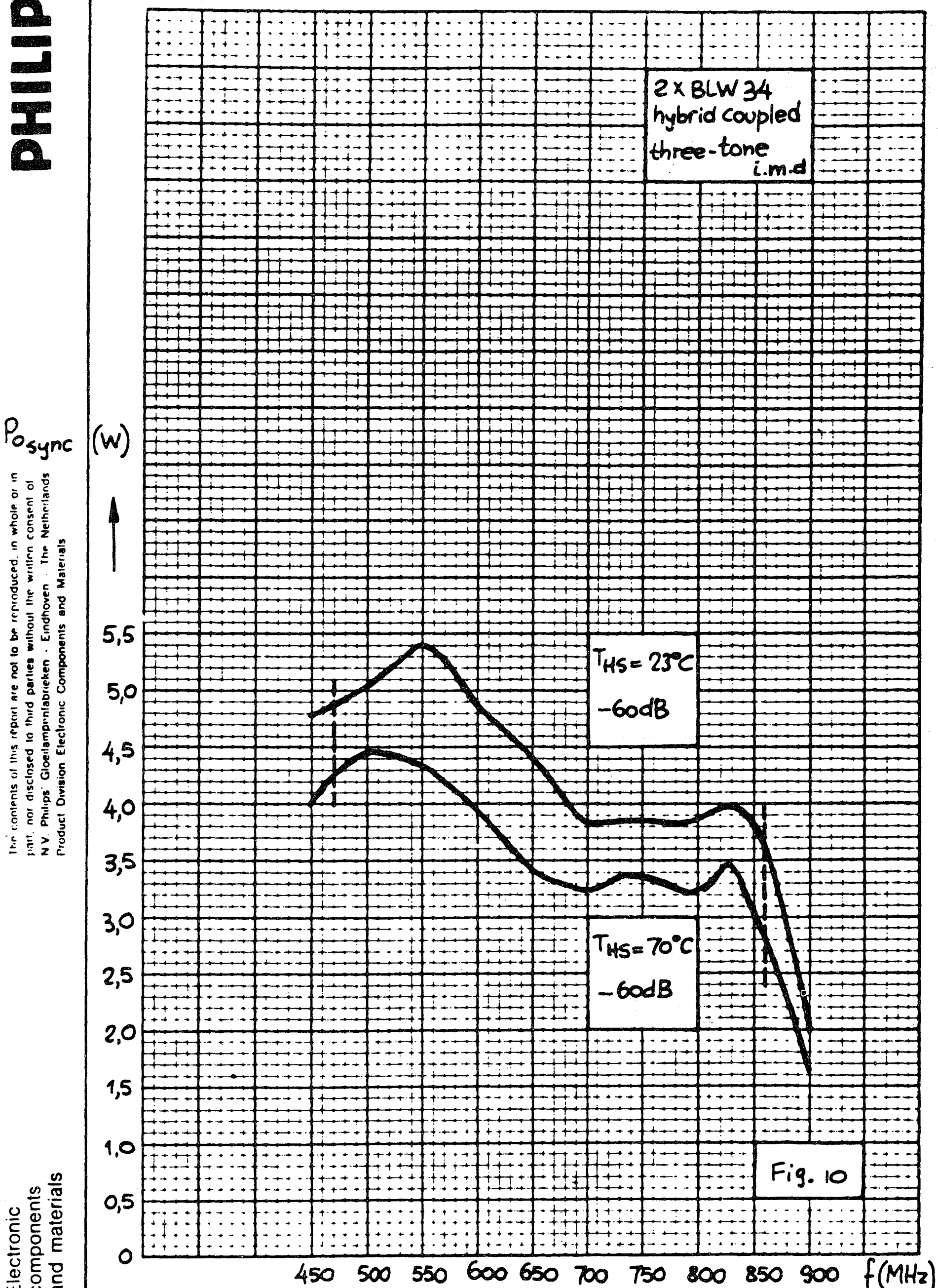

Fig. 10

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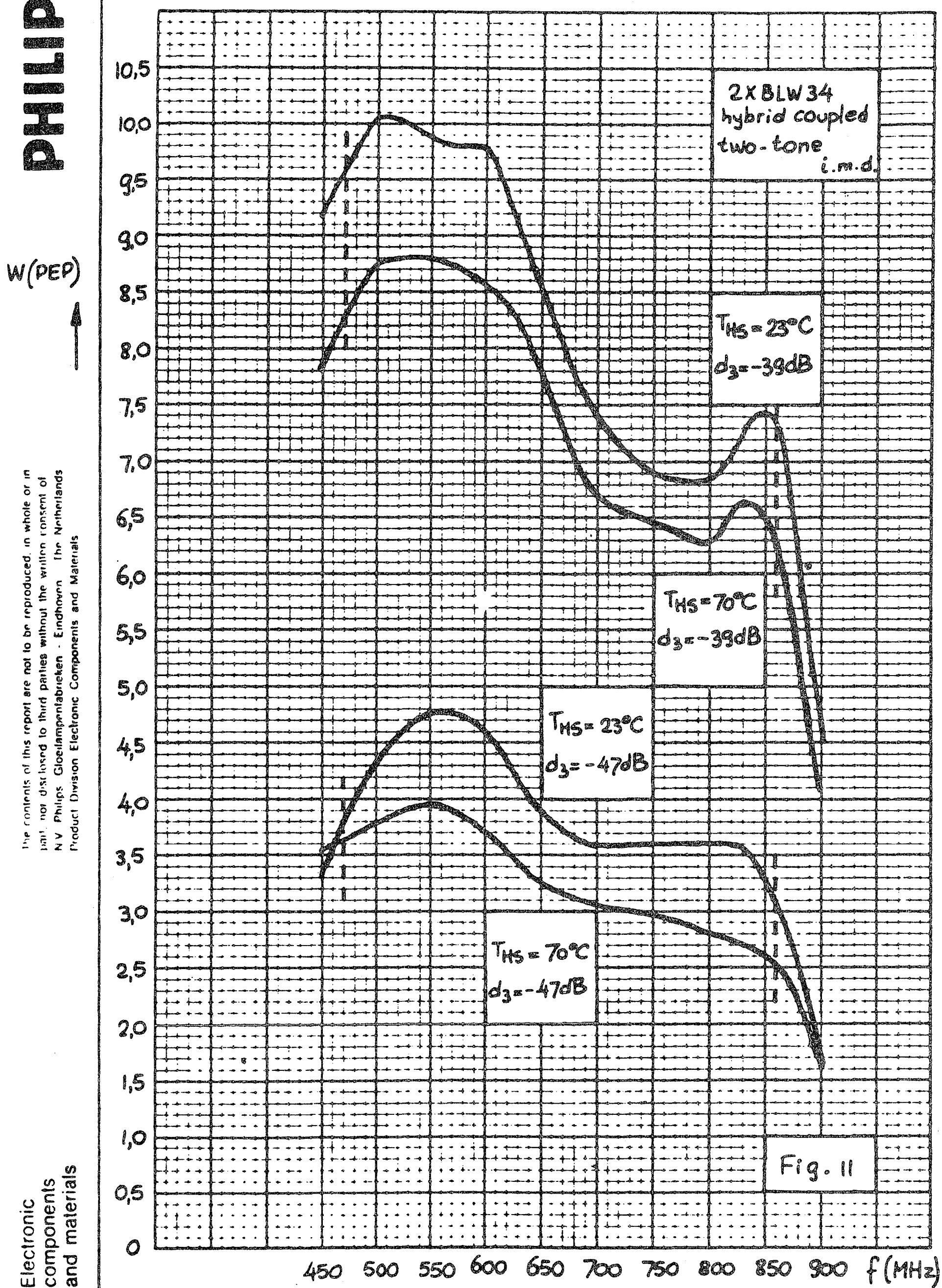

Fig. 11

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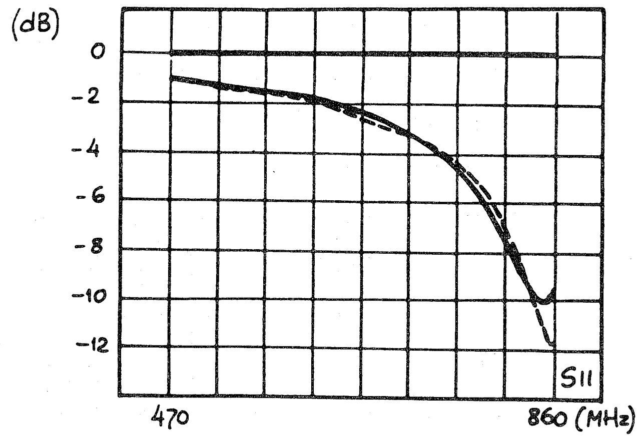
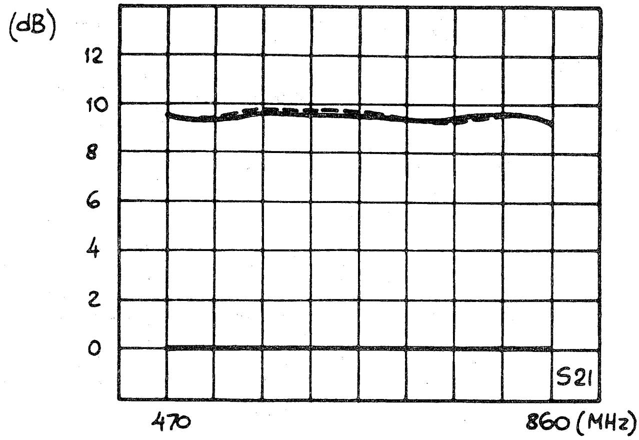


branch 1

--- branch 2

Fig. 12a



$T_{HS} = 70°C$

$V_{CE} = 25V$

$I_C = 600mA$

Fig. 12b

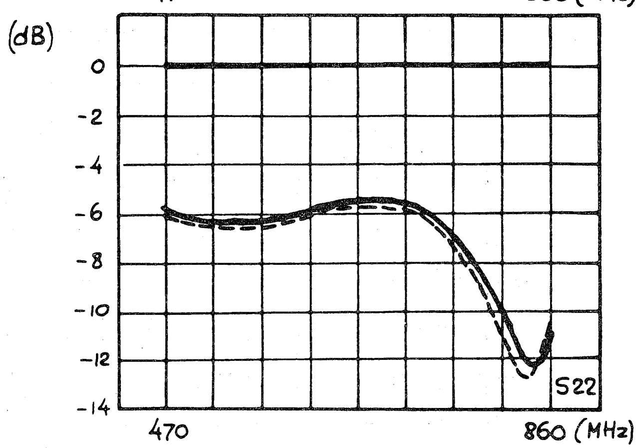


Fig. 12c

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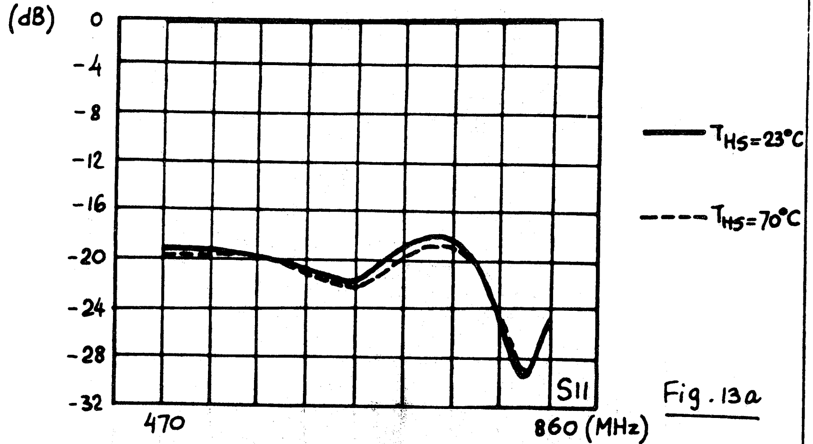

Fig. 13a

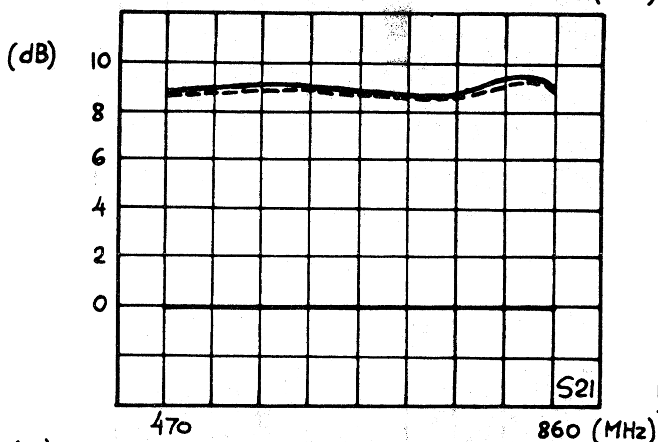

Fig. 13b

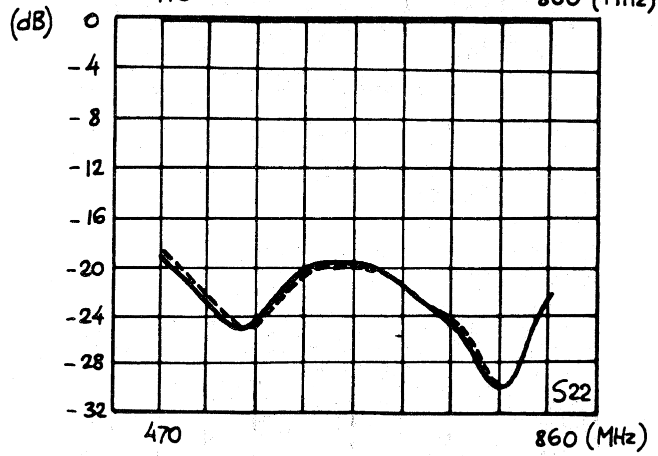

Fig. 13c

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laboratory report

number : EC07904 **date** : 21-12-1979

title : A WIDE-BAND CLASS-A LINEAR POWER AMPLIFIER (170-230 MHz) with two TRANSISTORS BLV33.

author : R.F.F.Zwanen

Electronic
components
and materials

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laboratory report

central application laboratory CAB
eindhoven - the netherlands

number: ECO 7904 date: 21-12-1979

project: 1064 pages: A1 ; R15 ;

title A WIDE-BAND CLASS-A LINEAR POWER AMPLIFIER (170-230 MHz) WITH 2 TRANSISTORS BLV 33

author R.F.F. Zwanen (Dev. Transm. And Microw. Dev.)

ABSTRACT

For application in driver or final stages of TV-transposers in band III (174-230MHz) a linear wideband power amplifier has been designed with 2 transistors BLV 33, coupled by means of 3dB-90^0 hybrids.
Each transistor is adjusted in class-A at V_{CE} = 25V and I_C = 3,25A.
A demonstration model showed a peak sync output power of 40W at a 3-tone I.M. distortion between -56 and -58dB. At this power level the cross-modulation varied from 6 to 7,5%.
The power gain is between 8,35 and 8,6 dB.

Appr. R.A. Pölzl

Advies Octrooi d.d. 1980-01-11	X AV	GV		B...		BL
Opgave Mamo d.d. 1980-01-17	X AV	X GV	X SP	B		BL
Datum: 21 dec. 1979	Mamo					

1. INTRODUCTION

For application in T.V. transposers and transmitters for band III a wideband linear power amplifier has been designed with 2 transistors BLV 33, coupled by means of 3dB - 90^{0} hybrids.
Each transistor is adjusted in class-A at V_{CE} = 25V and I_C = 3,25A.

2. Design of the amplifier

For class-A operation the BLV33 is specified at V_{CE}=25V, I_C = 3,25A. The corresponding typical gain, input and load impedance are given below:

Freq. (MHz)	Gain (dB)	Input Impedance (Ohm)	Load Impedance (Ohm)
174	11,3	0,68 + j1,20	2,70 + j1,19
202	10,1	0,68 + j1,43	2,30 + j0,87
230	9,08	0,68 + j1,64	1,99 + j0,52

A computer-aided circuit design, carried out by Mr. Hilbers (Central Application Laboratory) indicated a gain of 9,1dB $\pm$ 0,1dB and V.S.W.R. figures of 4,3 (170MHz), 2,78 (202MHz) and 1,18 at 230MHz for a single amplifier.
To obtain a high linear output and at the same time good input and output matching (V.S.W.R. $\leq$ 1,2) 3dB - 90^{0} hybrids are used. The reflected input power will be absorbed in the 50 Ohm resistor, matching the isolated port (see Fig. 1).
There are some small differences (value and place of chip capacitors) between the theoretical design and the practical circuit. For detailed information on computer-aided design see ref. 1-2-3.
Mainly due to the insertion loss of the 3dB hybrids the gain drops from 9dB to 8,5dB.

The transistors used in this particular amplifier are typical products, measured in a narrow band test amplifier and specified as follows:

V_{CE} = 25V - I_C=3,25A - Th=70^{0}C

Transistor type	:	BLV33	
Batch no.	:	MD 8-14 no 5	MD 8-14 no 10
Vision frequency	:	224,25MHz	
Output power (peak sync)	:	22,9W	22,6W
Intermodulation product	:	-55dB	-55dB
Gain	:	9,03dB	9,11dB

3. Adjustments of the amplifier

The amplifier consists of two equal BLV33 branches (see fig.1) and both transistors are seperately biased at V_{CE}=25V,I_C=3,25A. A schematic diagram and lay-out of the bias unit is given in fig. 2. Each branch was adjusted for maximum and flat gain by means of a high power sweep with a frequency range from 170 to 230MHz. The output power of the amplifier was leveled at 40W which means about 50% of the D.C. input power.
After that, both branches are coupled by means of 3dB - 90^0 hybrids, input and output matching of the complete amplifier are adjusted below a V.S.W.R. of 1,2 with the aid of capacitors C1-C2-C37-C38 as shown in fig. 1.

4. Assembling of the amplifier and mechanical data

Due to the dimensions of the p.c. board (220 x 210mm) 2 extruded blackened aluminium heatsiks (cat. no. 56293) are screwed together. The transistors are screwed on an aluminium plate (thickness 12mm) which on its turn is screwed on the heatsink. Special attention has been paid to the surface finishing to keep the thermal resistance as low as possible.
Fig. 3 showed the p.c. board and lay-out of the amplifier.
Dimensions: l=224mm - w=223mm - h=113mm. Weight: 7,5 kg.

A photograph of the amplifier is shown in fig.4.

5. Measured results

Frequency range	:	170 to 230 MHz			
Output power (peak sync)	:	30W	40W	50W	
I.M. distortion	:	-60dB	-56dB	-54dB	*
Cross modulation	:	3%	6/7,5%	10/15,5%	**
Output power for 1dB gain compression	:	100W			
Gain at 40W output level	:	8,5dB $\pm$ 0,1dB			
Input and output V.S.W.R.	:	$\leq$ 1,2			
Ambient temp.	:	25°C			
Heatsink temp.	:	65°C			
Transistor stud temp.	:	85°C			

In fig. 5 the typical results of crossmodulation and intermodulation measurements on a demonstration amplifier have been given. Figs. 6 and 7 shows the S-parameters of the complete amplifier.
The measuring set-up is depicted in fig. 8.

* Vision carrier -8dB, sound carrier -7dB, sideband signal -16dB, zero dB corresponds to peak sync level;
fsound=fvision + 5,5MHz, fsideband=fvision-1MHz to fvision + 6MHz

** Vision carrier 0dB, sound carrier -7dB: voltage variation of sound carrier (%) when the vision carrier is switched from -20dB to 0dB;fsound=fvision + 5,5MHz.

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6. Conclusion

Two transistors BLV33, coupled by means of 3dB - 90^{0} hybrids, can deliver an output power of typ. 40W with an associated gain of 8,5dB. Required D.C. input power approx. 165W.
Using a high power sweep with adjustable transistor output power leveling provides a suitable methode to adjust a linear wideband power amplifier.

7. References

Ref. 1: G.L. Matthaei - Tables of Chebyshev Impedance Transforming Networks of Low-Pass Filter Form. Proceedings of the IEEE August 1964, pp 939-963.

Ref. 2: A.H. Hilbers and M.J. Köppen - A wide-band linear power amplifier (470-860 MHz) with two transistors BLW34. C.A.B. report ECO 7901.

Ref. 3: A. Boekhoudt - Wideband linear power amplifier (174-230MHz) with two transistors BLV31 C.A.B. report ECO 7903.

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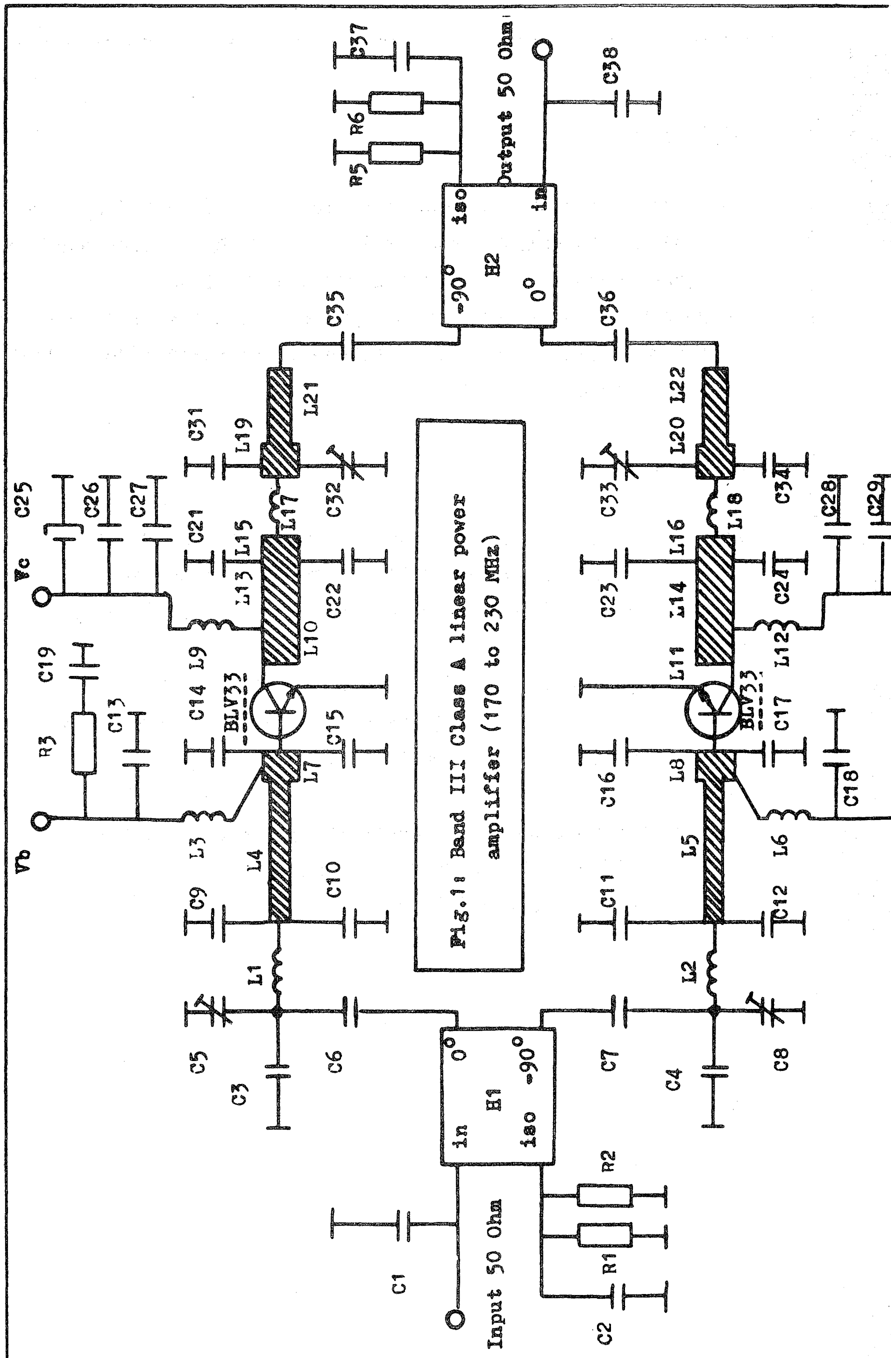

Fig.1: Band III Class A linear power amplifier (170 to 230 MHz)

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Parts list: Band III class A linear power amplifier (170 to 230MHz)

C1 = C38 = 1,5 pF, chip capacitor
C2 = C3 = C4 = C37 = 5,6 pF, chip capacitor
C5 = C8 = C32 = C33 = 1,8 to 10pF film dielectric trimmer (cat. no. 222280905002)
C6 = C7 = C35 = C36 = 220pF, chip capacitor
C9 = C10 = C11 = C12 = 18pF, chip capacitor
C13 = C18 = C27 = C28 = 1000 pF, chip capacitor
C14 = C15 = C16 = C17 = 100pF, chip capacitor
C19 = C20 = C26 = C29 = 330 nF metalized film capacitor (cat. no. 222235225334)
C21 = C23 = 68 pF, chip capacitor
C22 = C24 = 56 pF, chip capacitor
C25 = C30 = 10 uF (40V) electrolytic capacitor (cat. no. 222212117109)
C31 = C34 = 22 pF, chip capacitor
(chip capacitors: ATC type 100B - C - MSX - 500)
R1 = R2 = R5 = R6 = 100 Ohm, power metal film resistor, PR52 type (cat. no. 232219231001)
R3 = R4 = 10 Ohm, carbon resistor, CR68 type (cat. no. 23222141309)
H1 = H2 = 3dB - 90^0 coupler model no. 10262 -3, range 125 - 250 MHz, Anaren Microwave Inc.
L1 = L2 = 25nH; 2 turns enamelled Cu wire (1mm); int. diam. 5mm; length 5mm; leads 2 x 3 mm.
L3 = L6 = 90 nH; 5 turns closely wound enamelled Cu wire (1mm); int. diam. 4,5mm; leads 2 x 9mm
L4 = L5 = 60 Ohm stripline; w = 2mm; l = 30mm
L7 = L8 = 30 Ohm stripline; w = 6mm; l = 11mm
L9 = L12 = 20nH; Cu strip (1mm); l = 17mm; h = 5mm; w = 4mm
L10 = L11 = 30 Ohm stripline; w = 6mm; l = 8mm
L13 = L14 = 30 Ohm stripline; w = 6mm; l = 14mm
L15 = L16 = 30 Ohm stripline; w = 6mm; l = 4mm

L17 = L18 = 22nH; 2 turns closely wound Cu wire (1,5mm);
int. diam. 4,5mm; leads 2 x 3mm

L19 = L20 = 30 Ohm stripline; w = 6mm; l = 6mm

L21 = L22 = 50 Ohm stripline; w = 3mm; l = 15mm

The striplines are printed on double Cu-clad printed circuit board with epoxy fibre-glass dielectric (ε_r = 4,5); thickness 1/16 inch.

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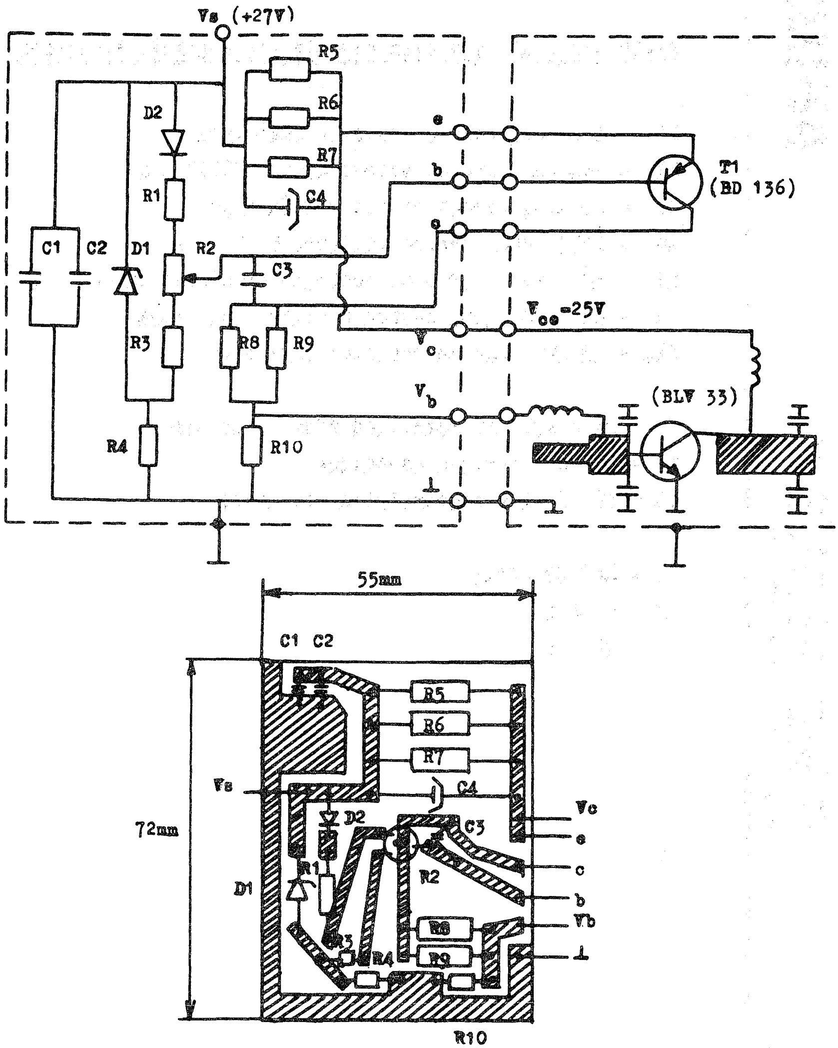

Fig.2 : Class A bias circuit for a single transistor BLV 33

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Parts list: Class A bias circuit for a single transistor BLV33

R1 = 150 Ohm, carbon resistor CR25 type
R2 = 100 Ohm, preset potentiometer CTP10 type
R3 = 10 Ohm, carbon resistor CR25 type
R4 = 1000 Ohm, carbon resistor CR25 type
R5 = R6 = R7 = 1,8 Ohm, rectangular wirewound resistor EH707 type
R8 = R9 = 180 Ohm, carbon resistor CR25 type
R10 = 33 Ohm, carbon resistor CR25 type

C1 = C3 = 100 nF, metalized film capacitor
C2 = 100 pF, ceramic capacitor
C4 = 10 uF, 40V electrolytic capacitor

D1 = BZY 88 (3V3)
D2 = BY 206
T1 = BD 136

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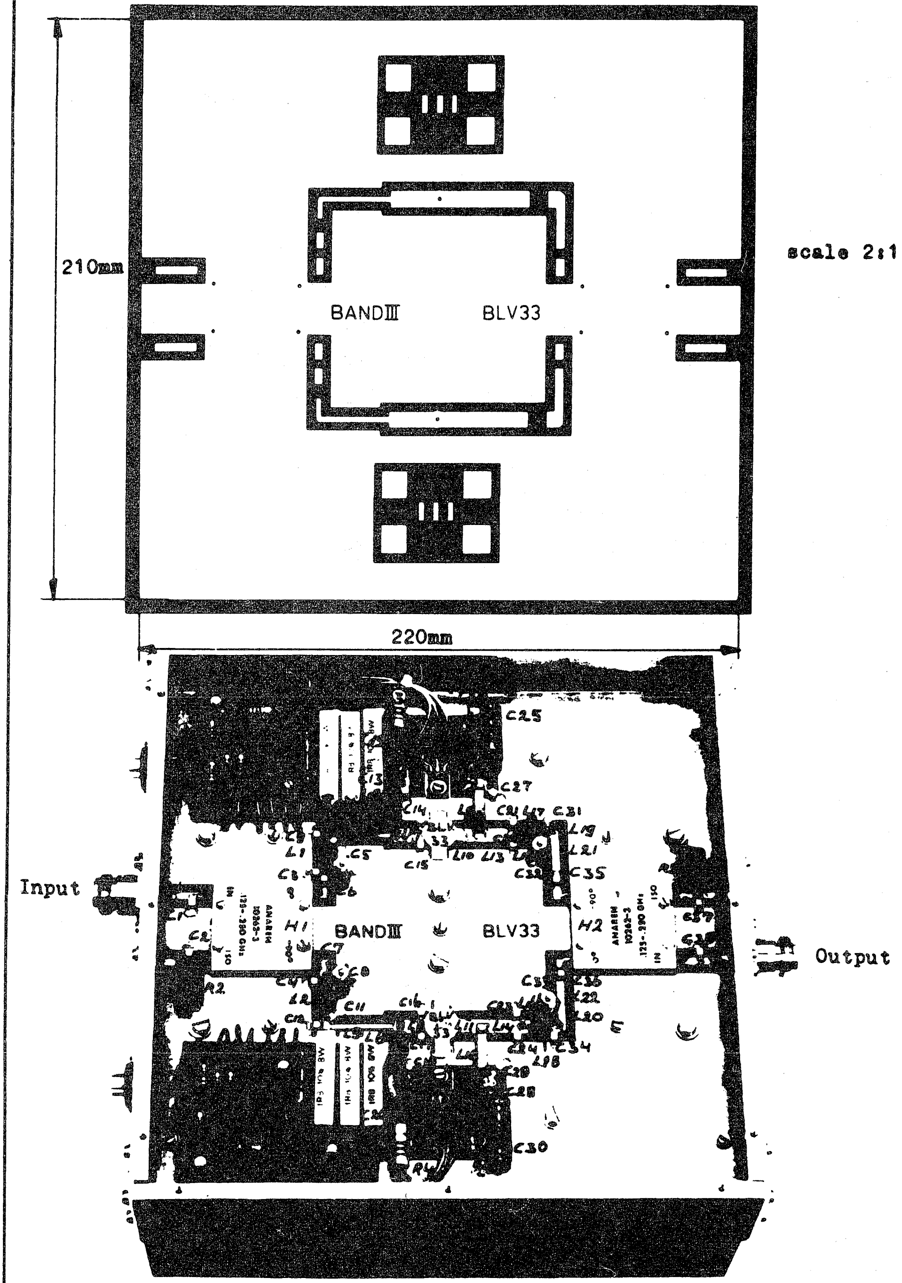

Fig.3 : P.C.Board and 2x BLV33 amplifier lay-out

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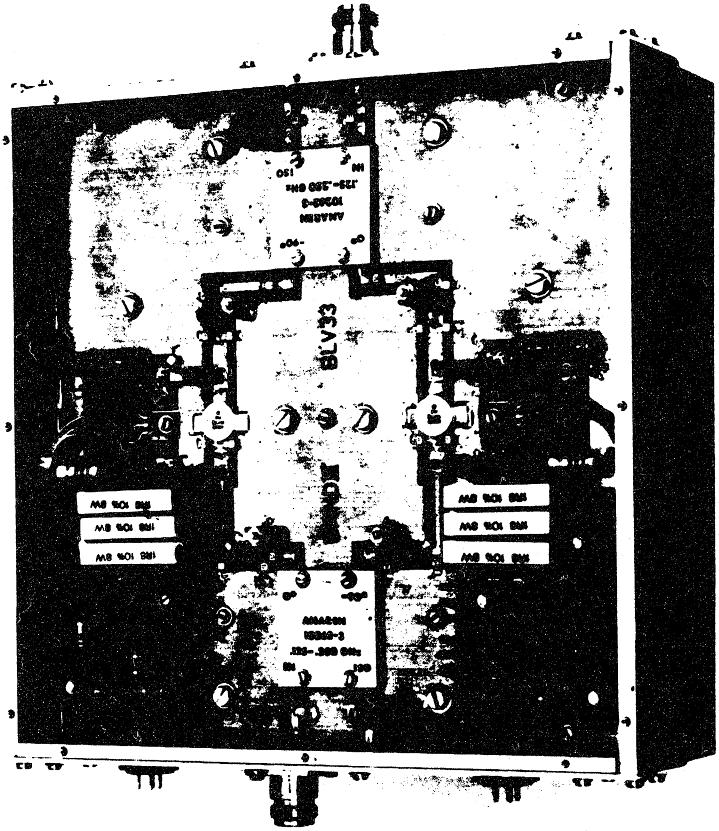

Fig. 4: Wide-band Class-A linear power amplifier (170 - 230 MHz) with two transistors BLV33

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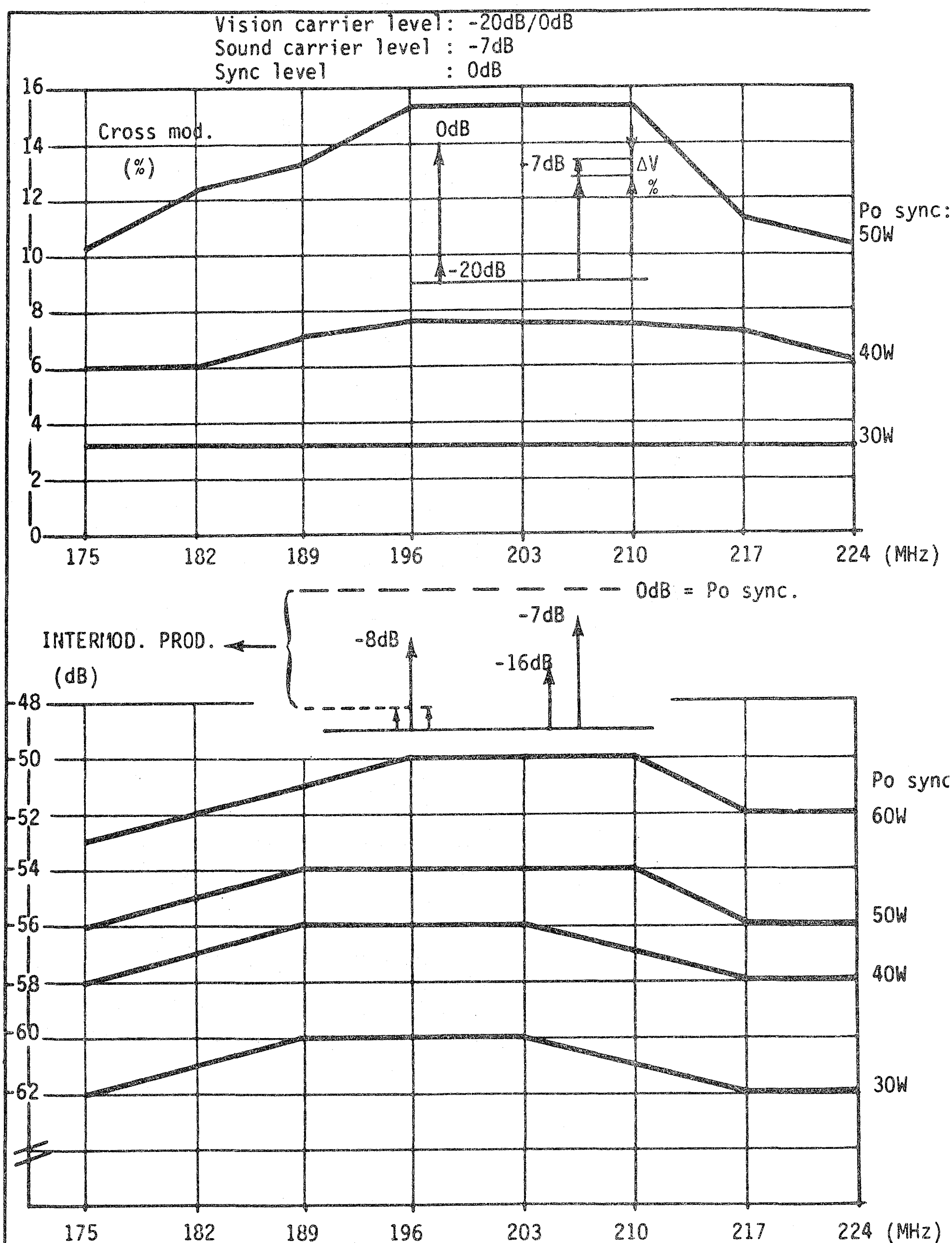


Fig. 5: Crossmod. and intermod. products of the 2 x BLV33 wideband Band III power amplifier.

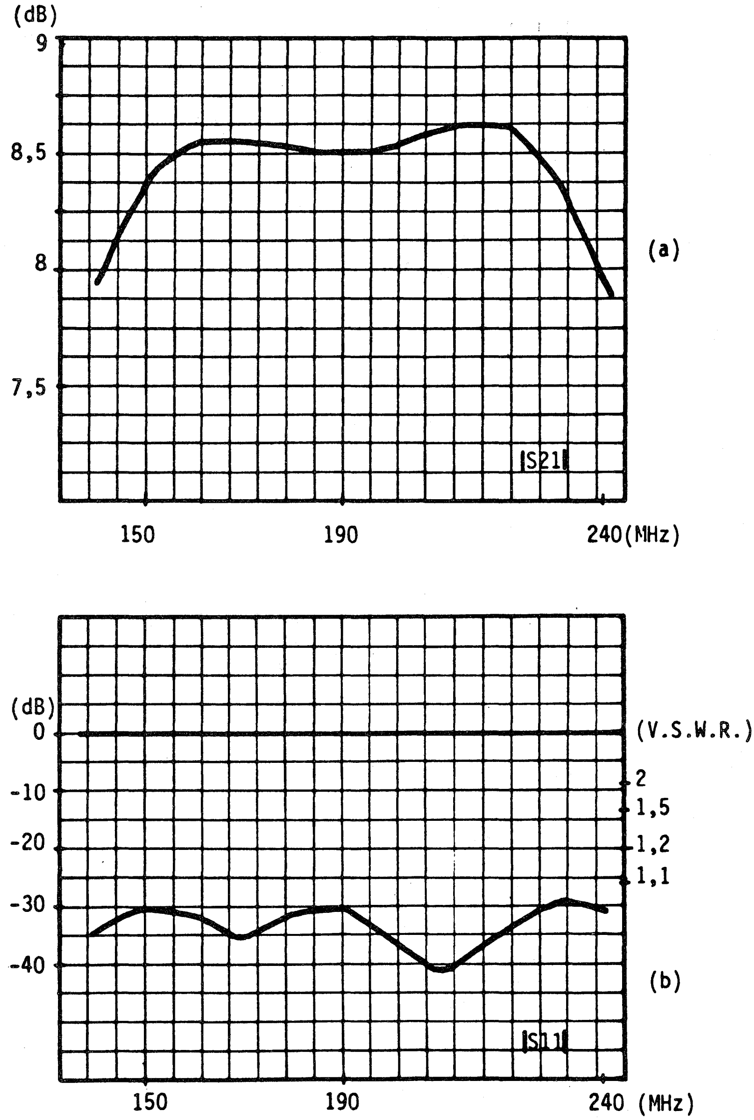
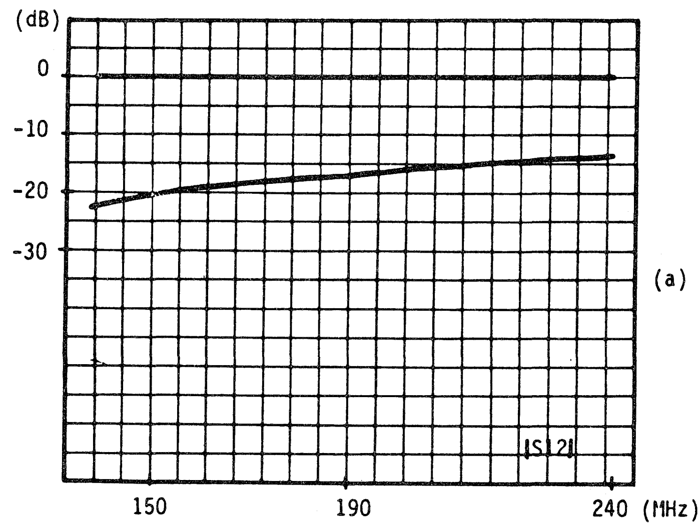


Fig. 6: 2 x BLV33 wideband Band III power amplifier:

a) Forward transducer gain

b) Input voltage standing wave ratio



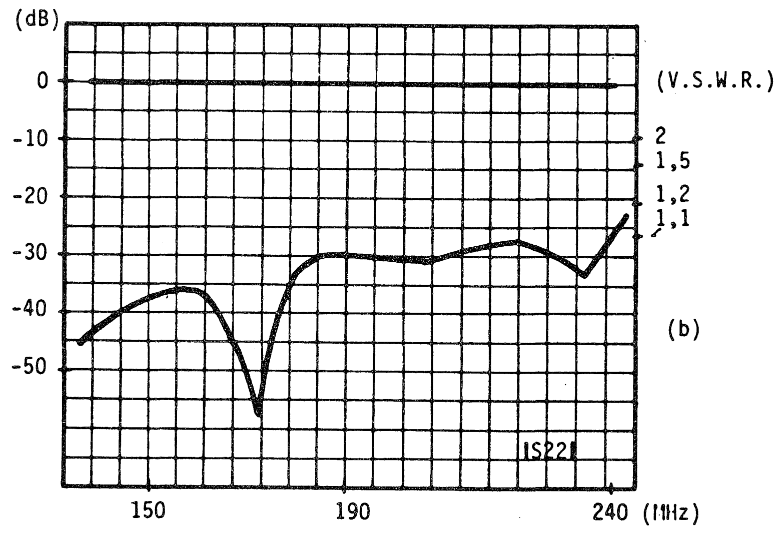


Fig. 7: 2 x BLV33 wideband Band III power amplifier:
a) Reverse transducer gain
b) Output voltage standing wave ratio

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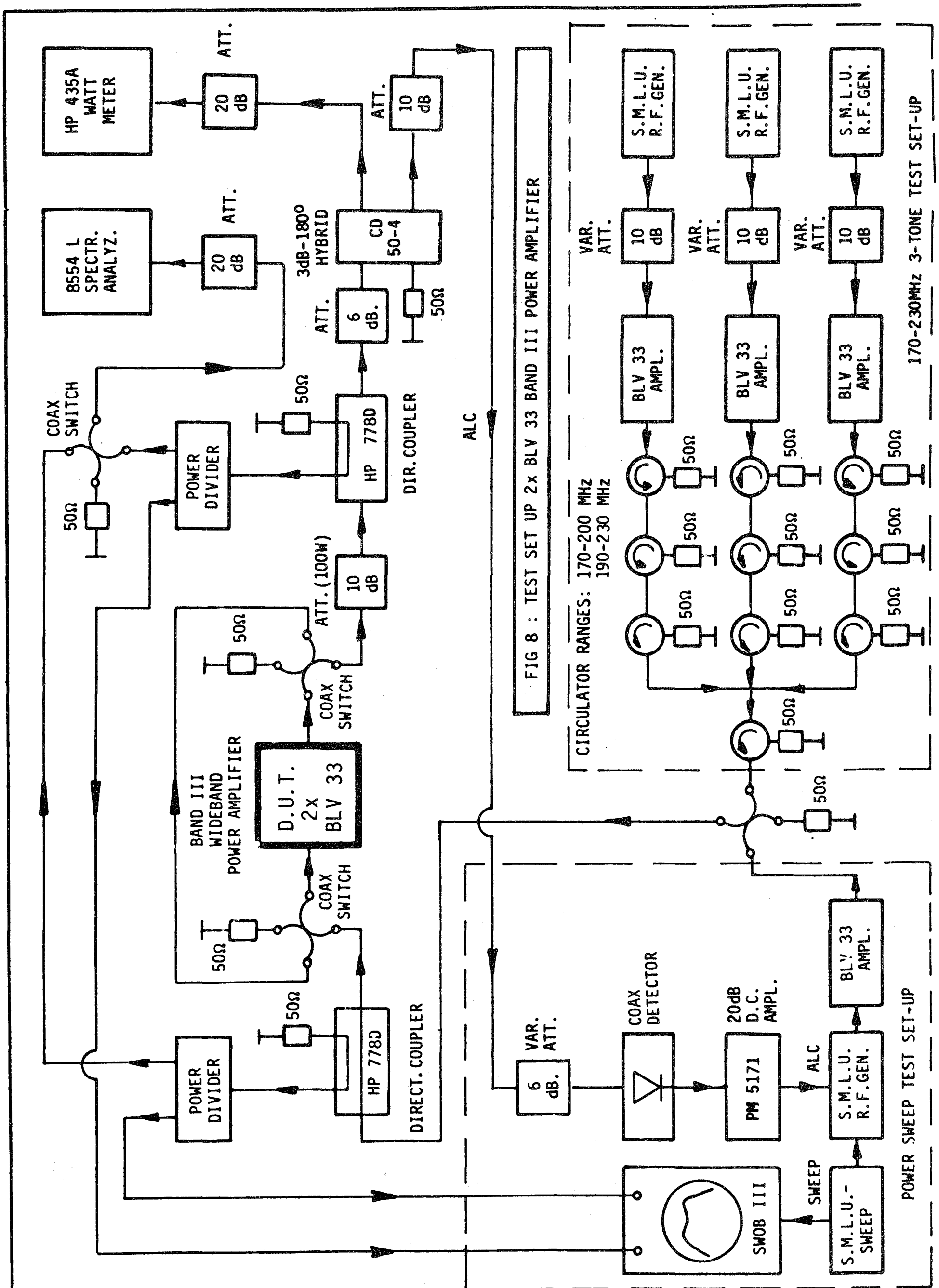

FIG 8 : TEST SET UP 2x BLV 33 BAND III POWER AMPLIFIER

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laboratory report

number : EC07905 **date** : 21-12-1979

title : WIDE-BAND LINEAR POWER AMPLIFIER (470-860 MHz) with two TRANSISTORS BLW98 (redesign of EC07704)

author : M.J.Köppen

Electronic components and materials

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laboratory report

central application laboratory CAB
eindhoven - the netherlands

number: ECO 7905 date: 21-12-1979

project: 6929 pages: A1 ; S6 ; R26

title WIDE-BAND LINEAR POWER AMPLIFIER (470-860 MHz) WITH TWO TRANSISTORS BLW98 (REDESIGN OF ECO 7704)

author M.J. Köppen

ABSTRACT

The TV transposer amplifier with 2 transistors BLW98 described in report ECO 7704 has been redesigned with the aim to increase the lower limit power gain by appr. 1dB and to reduce the size of the p.c. board. This has been achieved by increasing the number of matching sections and using thinner p.c. board. The transistors are operated in class-A at: V_{CE} = 25V, I_C = 850mA.
The power gain varies from 7,0 to 8,1dB. For a 3-tone I.M. distortion of -60dB the peak sync. output power is between 6,3 and 9,2W.
At a cross-modulation of 7% the output power ranges from 6,1 to 7,8W

Appr. R.A. Pölzl

Advies Octrooi d.d. 1980-01-11	X AV	GV		B		BL
Opgave Mamo d.d. 1980-01-17	X AV	X GV	X SP	B		BL
Datum: 21 dec 1979	Mamo					

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SUMMARY

Fig. A shows the basic circuit of the prototype with two transistors BLW98 both operating in class A.

The amplifier consists of two equal branches. Both are coupled at the input and output side by means of 3dB - 90^{0} coaxial hybrids (500-1000 MHz).

The circuit is mounted on a double-clad printed circuit board. To keep the losses in the transforming strip transmission lines sufficiently low the dielectric is PTFE glass fibre with an ε_r = 2,74 having a thickness of 1/32 inch. The fixed capacitors of the r.f. chain are of the ceramic multilayer chip type.

To obtain a high degree of linearity (-60dB i.m.d., three specified tones) the transistors have to operate in class A.
This class A operation is controlled by means of bias circuits with BD136 or BD138 (not shown in Fig. A).

The circuit contains all typical elements to assure stable and spurious-free operation even under mismatched input and output conditions.
The applied decoupling elements have been chosen so that the effect on the video step response is negligible.

This circuit is a redesign of the amplifier being published in C.A.B. report ECO 7704 (Ref. A). The target is $P_{o\ sync} \geq 6$ Watts and gain $\geq$ 7dB

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The theoretical design started with the output side in which the optimum load impedance (class A) varied between 9,98 + j7,31 ohms at 470 MHz and 4,55 + j4,33 at 860 MHz.

The output capacitance of the BLW98 is tuned out with the collector choke in the middle of the band.
Further, wide-band transformation to the 50 ohms amplifier load is arranged with the aid of an 8 elements transforming network.

On the transistor input the impedance ranges from 1,46 + j2,04 ohms at 470 MHz to 1,23 + j4,04 at 860 MHz, whilst the power gain varies with appr. 5dB/octave.

Reducing the gain variation based on permitting input mismatch at the lower frequencies has been done according to a method described in Ref. B.

The design has been theoretically optimized with the aid of a computer. For practical optimization and tuning the amplifier was inserted in a dynamic gain compression test set-up.

Figs. B1, 2, 3 show the small signal input and output reflection damping (S_{11} and S_{22}), whilst the power gain versus frequency is represented by the S_{21} curve. These results are given for a heatsink temperature of at least $60^{o}C$.

The output power $P_{o\ sync}$ versus frequency for three tone linearities of -60, -56 and -52dB is shown in Fig. C.

Special attention has been paid to the crossmodulation performance at several frequencies. The minimum $P_{o\ sync}$ for 7% crossmodulation amounts to 6,1 Watts (860 MHz).

As the curves show the targeted $P_{o\ sync} \geq 6$ Watts and gain $\geq$ 7dB could be reached.
The p.c. board area of the newly developed amplifier is about 60% of the original one.

Ref. A:

A.H. Hilbers and M.J. Köppen - Design of a wide band linear power amplifier (470-860 MHz) with two transistors BLW98.
C.A.B. report ECO 7704.

Ref. B:

O. Pitzalis. Jr and R.A. Gilson - Tables of impedance matching networks which approximate prescribed attenuation versus frequency slopes.
IEEE Transactions on microwave theory and techniques,
Vol. MTT-19, no. 4, April 1971, pp. 381-386.

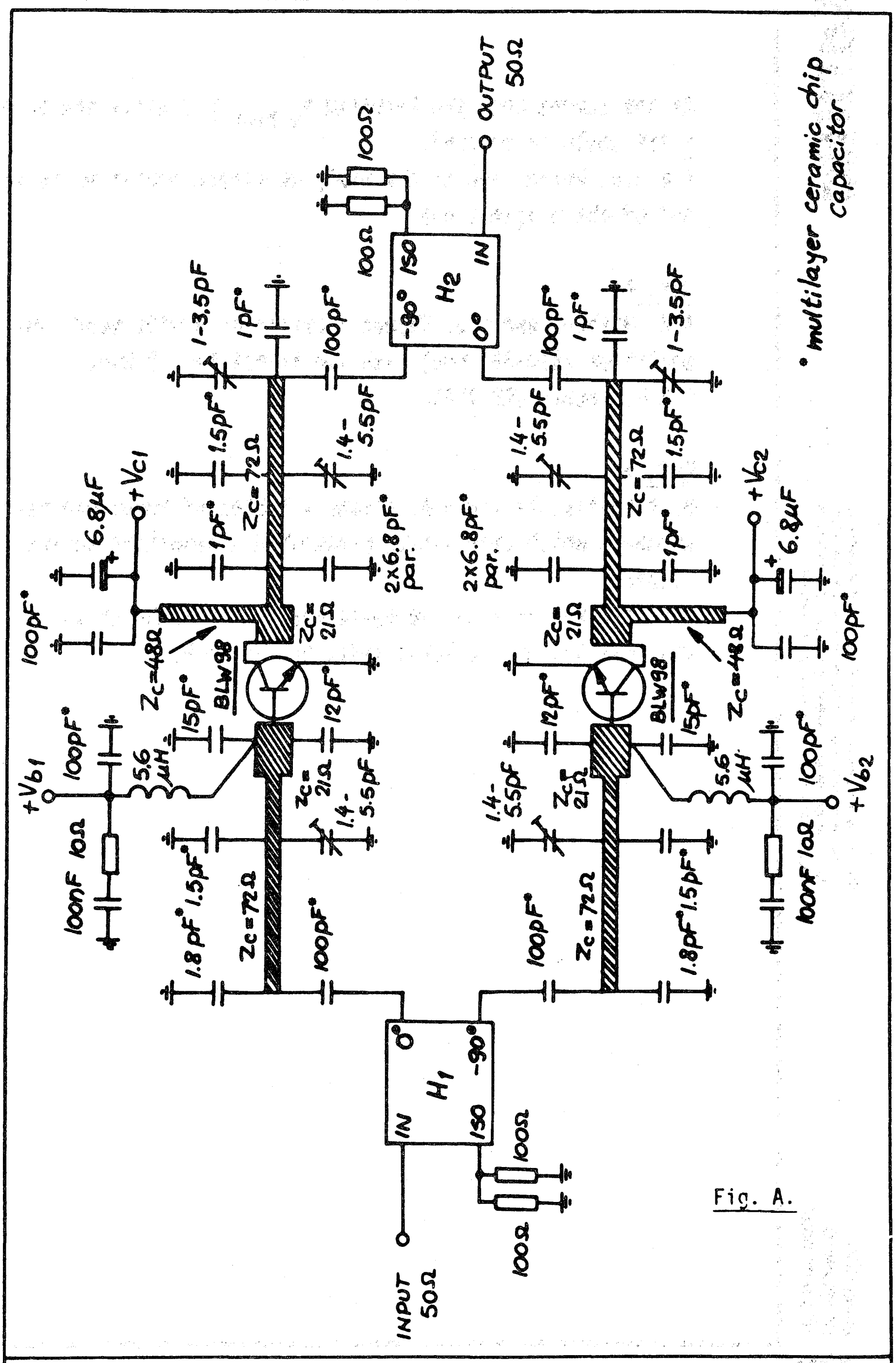

Fig. A.

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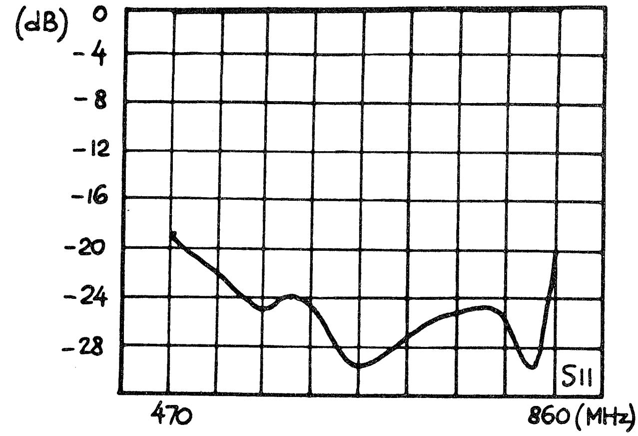

Fig. B1

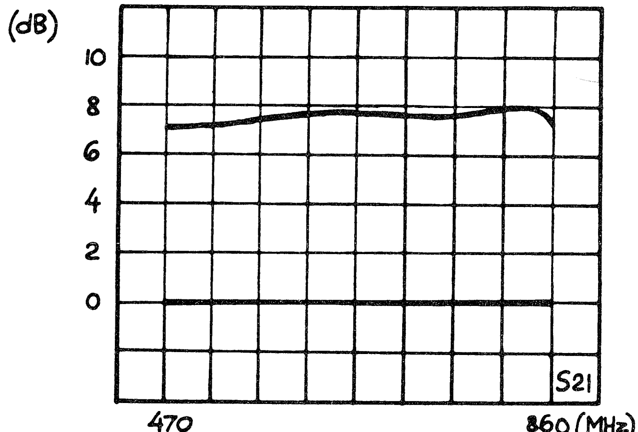

Fig. B2

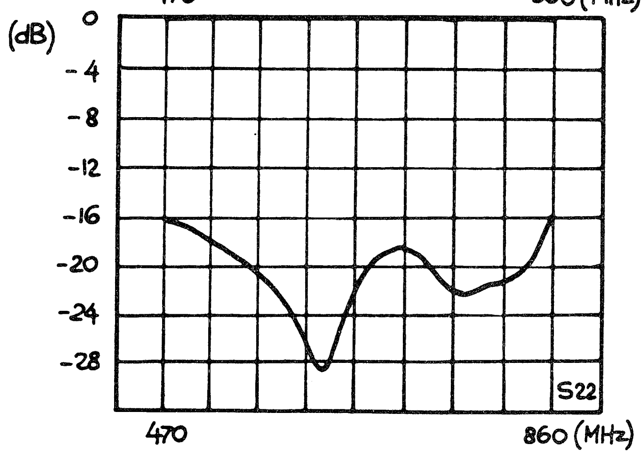

Fig. B3

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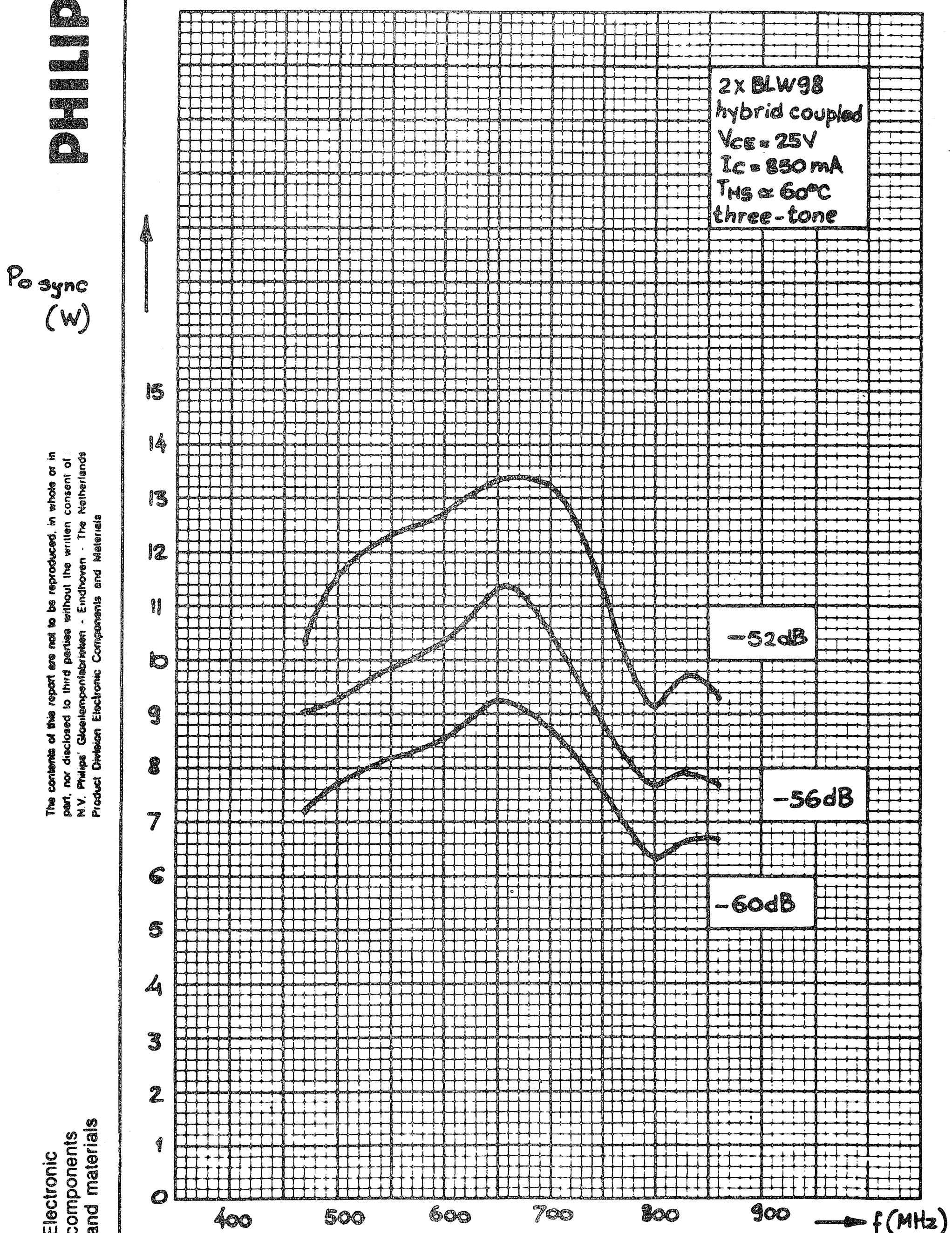

Fig. C

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1. INTRODUCTION

In 1977 a wide band linear VHF power amplifier equipped with two transistors BLW98 was developed.
The amplifier covered the TV bands IV-V (470-860 MHz) completely.

Application report ECO 7704 (Ref. 1) contains the description of the theoretical aspects and practical realization.

Since the power gain of the BLW98 declines with appr. 5dB/octave the power gain of the complete amplifier is kept constant within 1dB from 470-860 MHz by applying appropriate mismatch.
The resulting high VSWR especially at the lower frequency end is reduced to less than 1,6 by applying wide band 3dB-90^{0} coaxial hybrid couplers.

The quadrature configuration required for operating these couplers also facilitates the output power to be doubled by using a pair of BLW98 transistors.

Results from a two-tone test method with d_3 = -47dB showed that the amplifier was able to deliver 5,3 - 7,6 Watts PEP.

After two years we attempted to improve the performance of the amplifier by utilizing the experiences in developing a number of similar circuits and more advanced aligning methods.

The new target is $P_0 \geq 6$ Watts peak sync power for a gain of at least 7dB (was 6dB).
In contrast with the original application the tests should be done with three tones (-7, -8, -16dB) because this method delivers more reliable figures aggrevated to practical T.V. applications.

2. DESIGN CONSIDERATIONS (SINGLE AMPLIFIER)

2.1. Input and load impedances; gain

For class A operation the BLW98 is specified at V_{CE} = 25V and I_C = 850mA.

In the original amplifier the d.c. adjustment was changed to V_{CE} = 22,5V and I_C = 944mA. Such an operation point, requiring a lower load resistance to the transistor, resulted in an improved i.m.d. performance.

For the newly developed amplifier the 25V, 850mA operation point should be maintained, so the corresponding typical gain, input and load impedances according to the Data sheets are as follows:

f (MHz)	gain (dB)	$R_{i(series)}$ (Ohm)	$X_{i(series)}$ (Ohm)	$R_{L(series)}$ (Ohm)	$X_{L(series)}$ (Ohm)
470	11,2	1,46	+j 2,04	9,98	+j 7,31
547	10,1	1,43	+j 2,45	8,53	+j 6,96
636	9,0	1,38	+j 2,91	7,10	+j 6,34
739	7,9	1,31	+j 3,44	5,77	+j 5,46
860	6,9	1,23	+j 4,04	4,55	+j 4,33

The calculation of the circuit elements for the new amplifier is almost analogous to that of Ref. 1 and the amplifier with two pieces BLW34 being described in report ECO 7901 (Ref. 2).
To facilitate calculations an approximate equivalent circuit for the transistor input and output impedance can be given.

It is shown in Fig. 1.

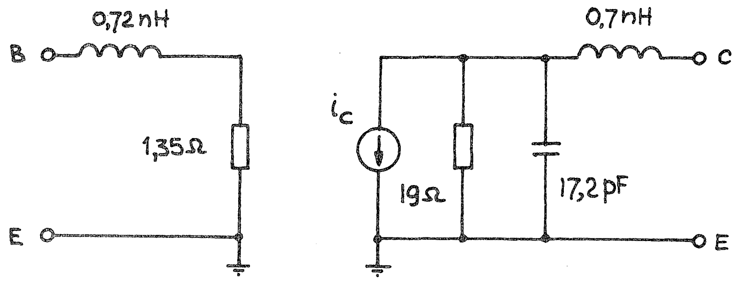

Fig. 1.

2.2. The output network

The circuit will be designed on double clad printed board with PTFE fibre glass at a dielectric having an ε_r = 2,74.

To minimize the dimensions of the p.c. board the strip transmission lines need to be smaller in width and somewhat shorter. So the circuit is designed now on p.c. board with a dielectric thickness of 1/32 inch instead of the 1/16 inch of the original amplifier.

The input and output networks start with a piece of stripline having a width of 6 mm, being the width of the base and collector leads. For 1/32 inch this means a characteristic resistance R_c of appr. 21 ohms.

The collector choke, tuning out the output capacitance of the transistor is executed as a stripline with a width of 2mm, corresponding to a characteristic resistance of appr. 48 ohms.

For practical reasons the choke is connected to the main transmission line at a distance of 3mm from the transistor edge.

There are some differences in the design procedure with respect to Ref. 1 and 2.

a. The length of the transistor collector lead has been cut off at 3 mm to decrease the parasitic capacitance at this point to a minimum.

b. An extra matching section has been added, what increases the number of elements from 7 to 8 of which 7 are optimized with a computer.

The results of the calculations before and after computer optimization are given in Fig. 2 and Table I.

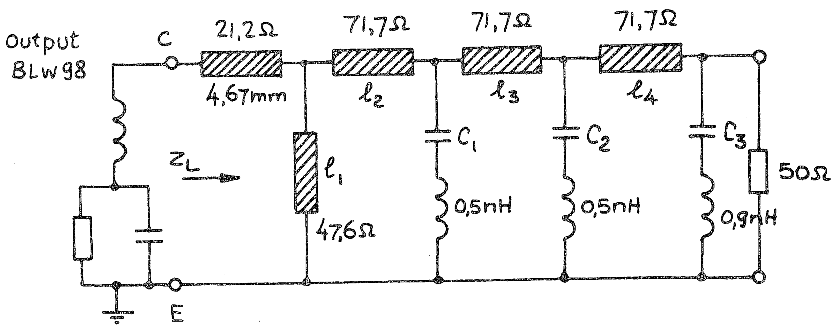

Fig. 2.

element	before optim.	after optim.	unit
l_1	21,7	20,5	mm
l_2	10,2	9,7	mm
C_1	15,18	13,92	pF
l_3	26,6	27,0	mm
C_2	9,63	6,65	pF
l_4	40,2	40,2	mm
C_3	3,47	2,23	pF
S_{max}	1,777	1,202	-

Table I

S_{max} is the maximum VSWR of the network.

The lengths given hold for air lines. The actual lengths on the p.c. board are shorter; the reduction factors are 1,455 for the 72 ohms lines, 1,492 for the 48 ohms line and 1,556 for the 21 ohms line.

The predicted minimum output power of the complete amplifier with two transistors BLW98 is:

$$P_{o2} = \frac{2.P_{o1}}{S_{max}} \cdot 0,95 = \frac{2 \cdot 3,5}{1,202} \cdot 0,95 = 5,53 \text{ Watts.}$$

P_{o1} is the minimum output power of one transistor in a narrow band circuit, S_{max} is as specified above and the factor 0,95 represents the power loss of the hybrid coupler at the output.

2.3. The input network

The design procedure is again almost equal to that given in Refs. 1 and 2. The differences are:

a. The length of the transistor base terminal has been reduced to 3mm, instead of the 1,7mm for the BLW34 transistor and 10mm of the original amplifier with BLW98.
The width is 6mm, corresponding with an R_c of appr. 21 ohms.

b. The R_c of the other series lines has been increased to appr. 72 ohms for the same reason as in the output network.

c. An extra matching section has been added, what increases the number of elements from 5 to 7 of which 5 are optimized with a computer.

d. To obtain a high power gain, both chip capacitors between base and emitter of the BLW98 are soldered as close as possible to the transistor envelope.

The results of the calculations are summarized in Fig. 3 and Table II.

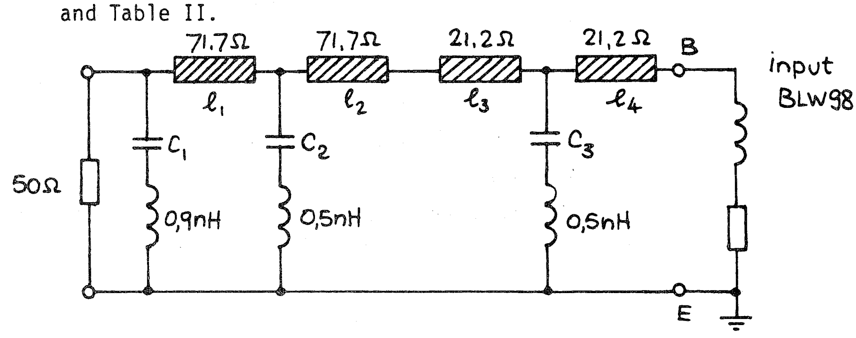

Fig. 3.

element	before optim.	after optim.	unit
C_1	1,24	1,89	pF
l_1	51,6	35,1	mm
C_2	4,42	7,66	pF
l_2	24,8	22,8	mm
l_3	3,11	3,11	mm
C_3	20,3	27,7	pF
l_4	1,56	1,56	mm
ΔG	$\pm$ 1,324	$\pm$ 0,084	dB

Table II*

ΔG is the resulting power gain variation caused by the transistor and the input network over the frequency band. The lengths of the lines hold for air as a dielectric. Transformation to striplines on a p.c. board is done in the same way as in the previous section.

The minimum power gain of the complete amplifier with two transistors BLW98 is expected to be:

$$G_o = G_T - 2A_H - A_1 - A_2$$

* Later it was discovered that during the initial calculation an error was made.
However this was corrected sufficiently by the optimization procedure.

in which: G_T = minimum power gain of BLW98 in a narrow band circuit.

A_H = power loss of one hybrid coupler

A_1 = reflection loss of input network

A_2 = reflection loss of output network

In practical figures this means:

$$G_0 = 6,5 - 2 \,.\, 0,2 - 0,04 - 0.08 = 5,98\text{dB}$$

The input VSWR of a single amplifier was calculated to vary from 8,97 to 470 MHz down to 1,32 at 860 MHz.

3. THE HYBRID COUPLED AMPLIFIER

3.1. Practical considerations

On previous pages the theoretical approach of the single amplifier was discussed.

In practice, it was the intention to realize a small compact amplifier on a p.c. board with input and output coaxial terminals (R_c = 50 ohms) in-line for easy cascading of several amplifiers.

Besides the wide-band properties it is the intention to obtain a higher output power, so two BLW98 branches are connected in parallel with the aid of two wide-band 3dB-90^0 coaxial hybrids on a 50 ohms basis.

So, by using coaxial hybrids the unacceptable situation that the amplifier shows a mismatch to the driver stage, has also been solved.

In such a configuration the output power will be nearly doubled; reduction of the input VSWR of the complete system to a value of around 1,2 may be explained from the properties of the hybrid coupler applied.
The reflected power will be absorbed in the resistor matching the isolated port. This resistor is 50 ohms and consists of two 100 ohms power metal film resistors in parallel.
The same has been done on the output side.

The p.c. board needs to be double copper clad with a PTFE glass fibre dielectric (ε_r = 2,74) for low losses at UHF.
The thickness is 1/32 inch.

Fig. 4 shows the circuit diagram of the complete 2 x BLW98 class A amplifier. The biasing circuit is drawn in Fig. 5.

The p.c. board is in Fig. 6 and the amplifier lay-out in Fig. 7.

For a correct earthing the upper earth sheet parts are connected to the lower sheet by soldering copper straps at the edges of the p.c. board. The black parts in Fig. 7 are the soldered copper straps.

The emitters are grounded as short as possible by applying copper straps under the emitter leads. For that reason the holes in the board are square instead of round.

Both transistors are screwed to an intermediate brass heatsink plate of 10mm thickness. The polished lower side of the p.c. board is in direct contact with the brass plate.
For this, small holes are drilled or fraised in the plate.

The coaxial connectors are of the SMA 50 ohms type, being soldered to upper and lower sheet and screwed to the intermediate heatsink.

3.2. Practical optimization

We started with optimization on a small signal basis with the circuit inserted in a network analyzer chain having swept S-parameter facilities.
So far, similar tuning methods have been applied as described in Refs. 2 and 3.

Because it is rather complicated to find the best compromise between an acceptable flat gain curve (S21) and sufficient output power with low i.m.d. a dynamic large signal optimization method has been applied.
Fig. 8 shows the block diagram.

This tuning method is based on the correlation between the single tone 1dB compression point and the i.m.d. figure of a linear amplifier.

In this set-up the swept output power level of the amplifier under test is kept constant and the required (detected) drive power is monitored on an oscilloscope screen (PM 3260).

The swept drive power is available from the sweep generator HP8620C in combination with RF plug-in unit HP86222A.
Because the output power of this system is too low viz. appr. 20mW (+13dBm) a combined amplifier with BLW32 and BLW33 (Ref. 3) cascaded with an amplifier with 2 x BLW34 (Ref. 2) have been added. The latter combination shows an overall gain of appr. 29dB in the range 470-860 MHz.

When the circuit under test is inserted in the chain of Fig. 8 the input power measured on port C of hybrid 1 corresponds in principle with the gain curve (S_{21}) being measured with the network analyzer; in fact one is the inverse of the other.

When the drive level is slowly increased, the shape of the gain curve changes somewhat when compression starts.
By careful retuning of the amplifier the shape of the gain curve can be corrected again in the direction of the original smaller signal curve.

In this way the amplifier of Ref. 2 was aligned.

Fig. 8 shows that a PM5715 pulse generator is added to the chain. It creates an amplitude variation of the sweep generator (via the external AM input). So, the compression level can be monitored in a dynamical way, what improves the alignment procedure.

The actual single tone output power has been measured with the aid of the calorimetric watt meter HP 435A when the action of the sweep is stopped and the external AM switched off.
Fig. 9 gives an idea of the screen picture with and without compression visible.

3.3. Intermodulation, VSWR and gain measurements

For i.m.d. measurements on television systems the post offices advise and apply the 3-tone test method (vision carrier -8dB, sound carrier -7dB , side band signal -16dB; zero dB corresponds to peak sync level). According to the CCIR system the frequency is given by the vision carrier and the sound carrier is 5,5MHz above this one.

The wide-band test set-up is according to the chain in report ECO 7901, page R21, Fig. 9 (Ref. 2).

In this set-up first the sound and vision carriers are each buffered by a circulator and then joined in a wide-band coaxial hybrid.
Then the sideband signal from the (smaller) generator SMLU is added in a second hybrid (no circulator applied).

The complete three-tone signal passes a low-pass filter (700 or 1000MHz cut-off depending on the input frequency), a continuously variable attenuator and a circulator (three different types are needed to cover at least the range 470-860MHz).

The output power is measured with a 435A calorimetric watt meter (multiplication factor for three-tone -7, -8, -16dB operation is 2,61) and the i.m.d. observed via a loosely coupled 50 ohms pick-off device with a HP 8558B spectrum analyzer.

Fig. 10 shows the three-tone i.m.d. results measured on a complete hybrid coupled 2 x BLW98 amplifier.
$P_{o\ sync}$ was measured for i.m.d. distances of -52, -56 and -60dB.
The heatsink temperature amounts to appr. 60°C.

Finally, input and output VSWR and gain figures were measured only under small signal conditions. The VSWR figures of the input and output are expressed in reflection damping (resp. S_{11} and S_{22}) on a 50 ohms basis (see Figs. 11, 12, 13).

According to Fig. 11 the minimum reflection damping of the input of the coaxial hybrid amounts to S_{11} = -19dB at 470 MHz what corresponds with a maximum VSWR of 1,25.

On the output side the worst reflection damping amounts to S_{22} = -16dB (VSWR = 1,38) at both band ends.
Both values may be mainly explained from the specified maximum VSWR = 1,25 of the applied Anaren hybrids and the fact that the load of the isolated port consists of two metal film power resistors of 100 ohms in parallel having low frequency tolerances of 5%.

The power gain, represented by S_{21}, is at least 7dB (Fig. 12).
For a second check this gain curve has been measured step by step for an output power of 600mW.
Fig. 13 shows the results.

3.4. Cross-modulation

There is a trend now that the cross modulation behaviour of a linear amplifier becomes more important.
For this reason tests have been done and the results given in this report (Figs. 15 to 19).

These tests could be done in the 3-tone chain with the sideband carrier switched off.
The vision carrier has been made equal to $P_{o\ sync}$ (0dB), instead of -8dB, whilst the sound carrier remains -7dB (5 x power).

By means of a coaxial switch and a 20dB attenuator, inserted in the vision carrier chain, this carrier can be switched from 0 to -20dB.

Increasing both tones with the same factor and switching periodically the vision carrier from 0 to -20dB the procentual influence on the sound carrier (with the spectrum analyzer switched to linear operation) is measured with respect to its original level.

4. CONCLUSIONS

On preceding pages the theoretical and practical design has been described of a wide-band (470-860MHz) high quality linear amplifier being equipped with two BLW98 transistors operating in class A. This amplifier is a modified version of that already published in report ECO 7704 (Ref. 1).
The target $P_{o\ sync} \geq 6$ Watts (was ≥ 6 Watts) and power gain ≥ 7dB (was ≥ 6dB) was reached, whilst the area of the new p.c. board is only about 60% of the original one.

There are some small differences between the theoretical design and the practical circuit, viz.

- The calculated value for the chip capacitors $C_7 = C_{10}$ was 3,9pF. In practice 1,5pF appeared to be a better choice.

- $C_{22} = C_{24} = 6{,}8$pF was too small, so $C_{21} = C_{23} = 1$pF has been added.

- $C_{29} = C_{32}$ was planned as a fixed chip capacitor of 1,2pF; tuning this point with a variable capacitor of 1-3,5 pF gave advantages.

PHILIPS

5. REFERENCES

Ref. 1: A.H. Hilbers and M.J. Köppen - Design of a wide-band linear power amplifier (470-860 MHz) with two transistors BLW98. C.A.B. report ECO 7704.

Ref. 2: A.H. Hilbers and M.J. Köppen - A wide-band linear power amplifier (470-860 MHz) with two transistors BLW34. C.A.B. report ECO 7901.

Ref. 3: A.H. Hilbers and M.J. Köppen - Wide-band linear power amplifiers (470-860 MHz) with the transistors BLW32 and BLW33. C.A.B. report ECO 7806.

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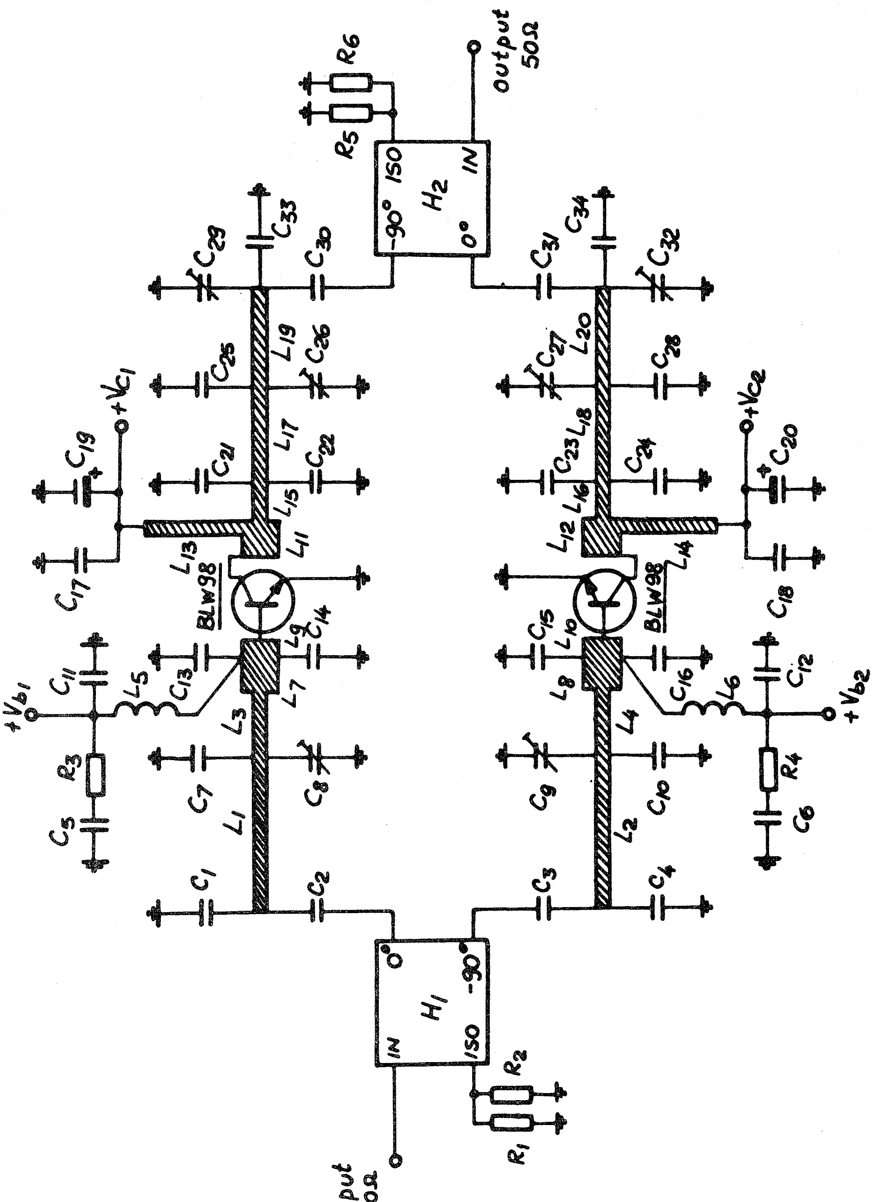

Fig. 4.

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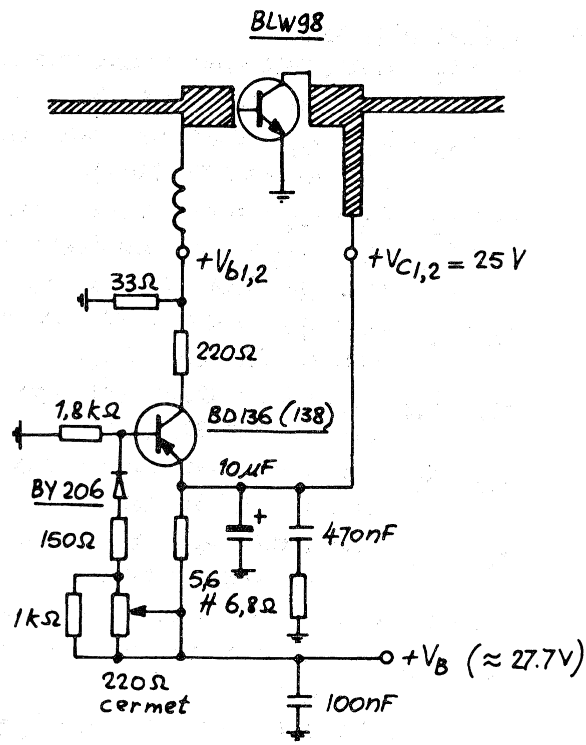
BLW98
+$V_{b1,2}$
+$V_{C1,2}$ = 25 V
33Ω
220Ω
1,8 kΩ
BD136 (138)
BY 206
10 μF
150Ω
470nF
5,6 // 6,8Ω
1 kΩ
+V_B (≈ 27.7V)
220Ω cermet
100nF

Fig. 5.

6. LIST OF COMPONENTS

$C_1 = C_4$ = 1,8pF, multilayer ceramic chip capacitor
(ATC type: 100A - 1R8 - B - P_x - 50)

$C_2 = C_3 = C_{30} = C_{31}$ = 100pF, multilayer ceramic chip capacitor
(cat. no. 2222 851 13101).

$C_5 = C_6 = C_{39} = C_{40}$ = 100nF, polyester capacitor.

$C_7 = C_{10} = C_{25} = C_{28}$ = 1,5pF, multilayer ceramic chip capacitor
(ATC type: 100A - 1R5 - B - P_x - 50).

$C_8 = C_9 = C_{26} = C_{27}$ = 1,4 to 5,5pF film dielectric trimmer
(cat. no. 2222 809 09001).

$C_{11} = C_{12} = C_{17} = C_{18}$ = 100pF, multilayer ceramic chip capacitor
(cat. no. 2222 852 13101).

$C_{13} = C_{16}$ = 15pF, multilayer ceramic chip capacitor
(cat. no. 2222 851 13159).

$C_{14} = C_{15}$ = 12pF, multilayer ceramic chip capacitor
(cat. no. 2222 851 13129).

$C_{19} = C_{20}$ = 6,8µF, 63V, electrolytic capacitor.

$C_{21} = C_{24}$ = 2 x 6,8pF in parallel, multilayer ceramic chip capacitors
(ATC type: 100A - 6R8 - J - P_x - 50).

$C_{22} = C_{23} = C_{33} = C_{34}$ = 1pF, multilayer ceramic chip capacitor
(ATC type: 100A - 1R0 - B - P_x - 50).

$C_{29} = C_{32}$ = 1 - 3,5pF film dielectric trimmer
(cat. no. 2222 809 05001).

$C_{35} = C_{36}$ = 10µF, 63V, electrolytic capacitor.

$C_{37} = C_{38}$ = 470nF, polyester capacitor.

$R_1 = R_2 = R_5 = R_6$ = 100Ω (± 5%), power metal film resistor
PR37 type (cat. no. 2322 191 31001).

$R_3 = R_4 = R_{21} = R_{22}$ = 10Ω (± 5%) carbon resistor; CR25 type.

$R_7 = R_8$ = 1,8kΩ (± 5%); CR25 type.

$R_9 = R_{10}$ = 1kΩ (± 5%); CR25 type.

$R_{11} = R_{12}$ = 33Ω (± 5%); CR25 type.

$R_{13} = R_{14}$ = 150Ω (± 5%); CR25 type.
$R_{15} = R_{16}$ = 220Ω, cermet preset potentiometer
$R_{17} = R_{18}$ = 220Ω (± 5%); power metal film resistor
PR52 type (cat. no. 2322 192 32201).
$R_{19} = R_{20}$ = 5,6Ω (± 5%) and 6,8Ω (± 5%) in parallel;
enamelled wire-wound resistors WR 0617 style.

$H_1 - H_2$ = ultra-miniature 3dB-90° coupler
model no. 10264-3, range 0,5 - 1,0 GHz.
Anaren Microwave Inc.
$L_1 = L_2$ = stripline (Z_c = 72Ω), 24,1 x 1,0 mm² *
$L_3 = L_4$ = stripline (Z_c = 72Ω), 15,7 x 1,0 mm² *
$L_5 = L_6$ = 5,6μH; 29 turns closely wound enamelled Cu wire
(0,2 mm); int. dia 3,0 mm.
$L_7 = L_8$ = stripline (Z_c = 21Ω), 2,0 x 6,0 mm² *
$L_9 = L_{10}$ = stripline (Z_c = 21Ω), 1,0 x 6,0 mm² *
$L_{11} = L_{12}$ = stripline (Z_c = 21Ω), 3,0 x 6,0 mm² *
$L_{13} = L_{14}$ = stripline (Z_c = 48Ω), 13,8 x 2,0 mm² *
$L_{15} = L_{16}$ = stripline (Z_c = 72Ω), 6,7 x 1,0 mm² *
$L_{17} = L_{18}$ = stripline (Z_c = 72Ω), 18,5 x 1,0 mm² *
$L_{19} = L_{20}$ = stripline (Z_c = 72Ω), 27,6 x 1,0 mm² *

* These striplines are printed on double Cu-clad printed circuit board with PTFE fibre-glass dielectric (ε_r = 2,74); thickness 1/32"

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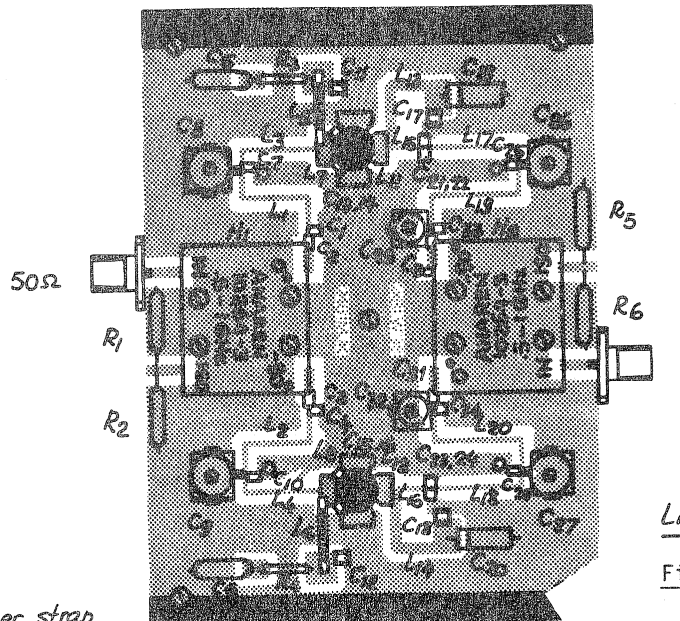
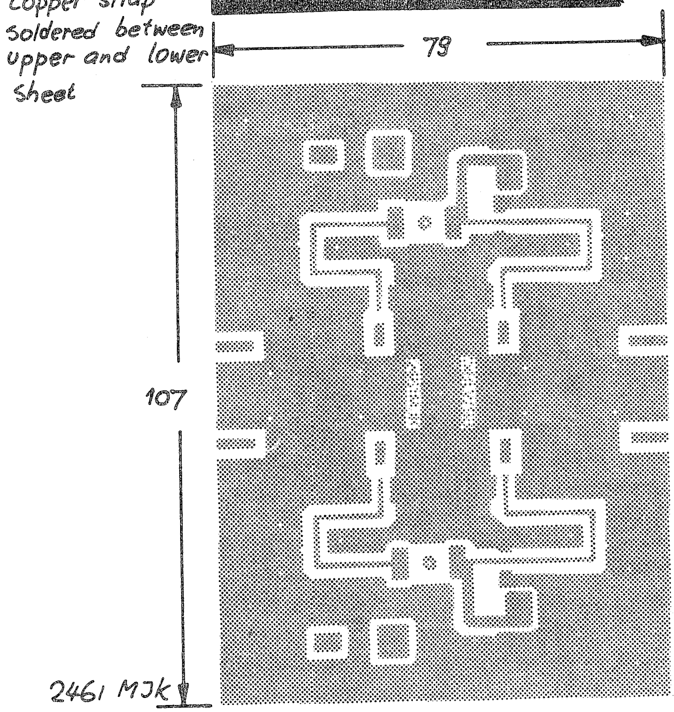


LAY-OUT

Fig. 7.



P.C. BOARD

1/32 inch PTFE double Cu clad

Fig. 6.

8222 031 76361

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DYNAMIC COMPRESSION

TEST SET-UP

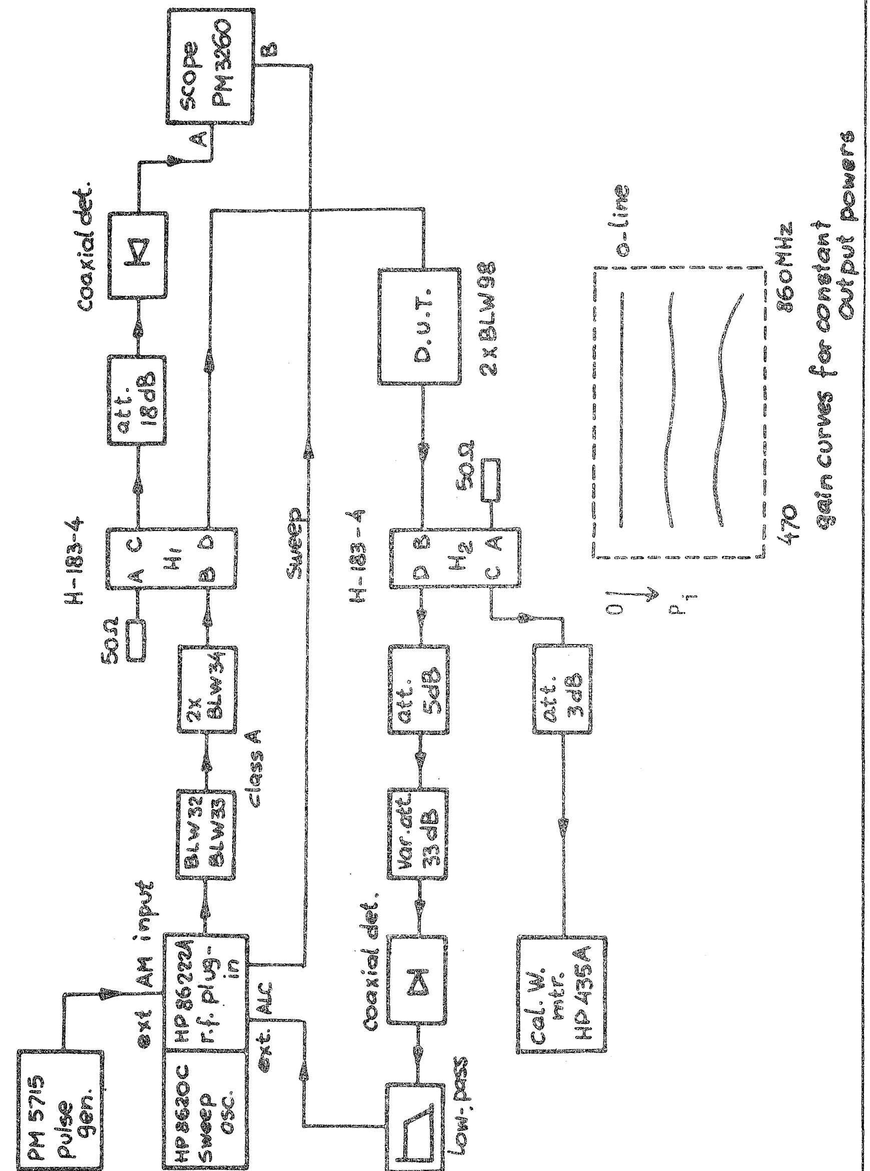

Fig. 8.

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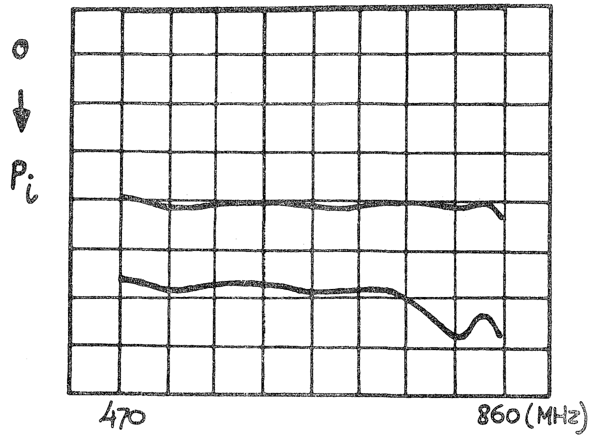

Fig. 9.

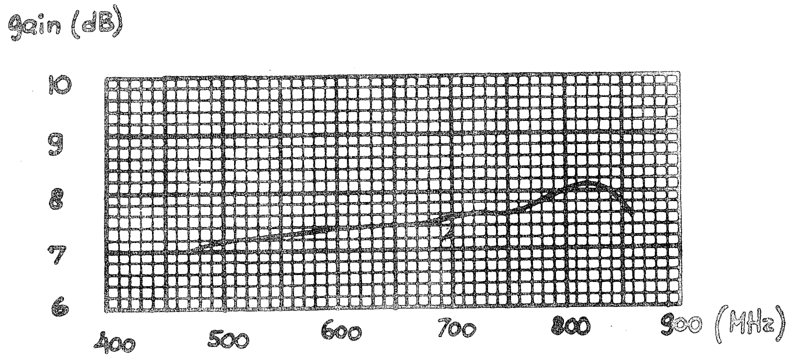

Fig. 14.

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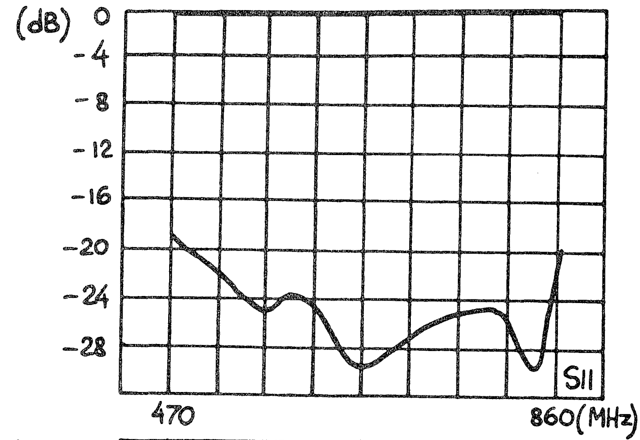

Fig. 11.

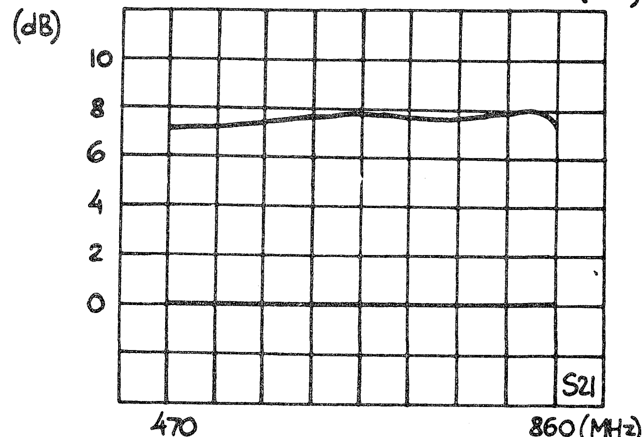

Fig. 12.

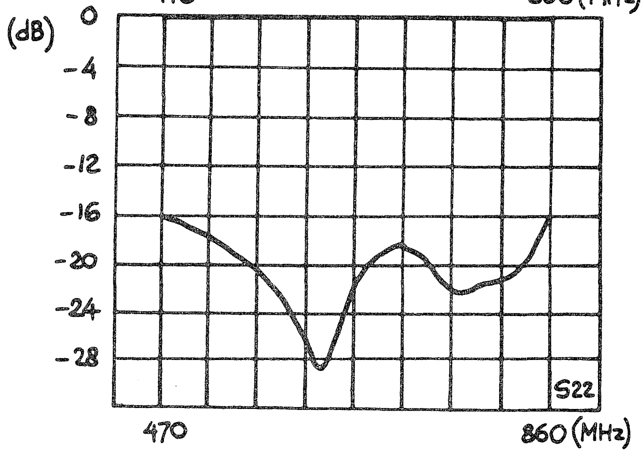

Fig. 13.

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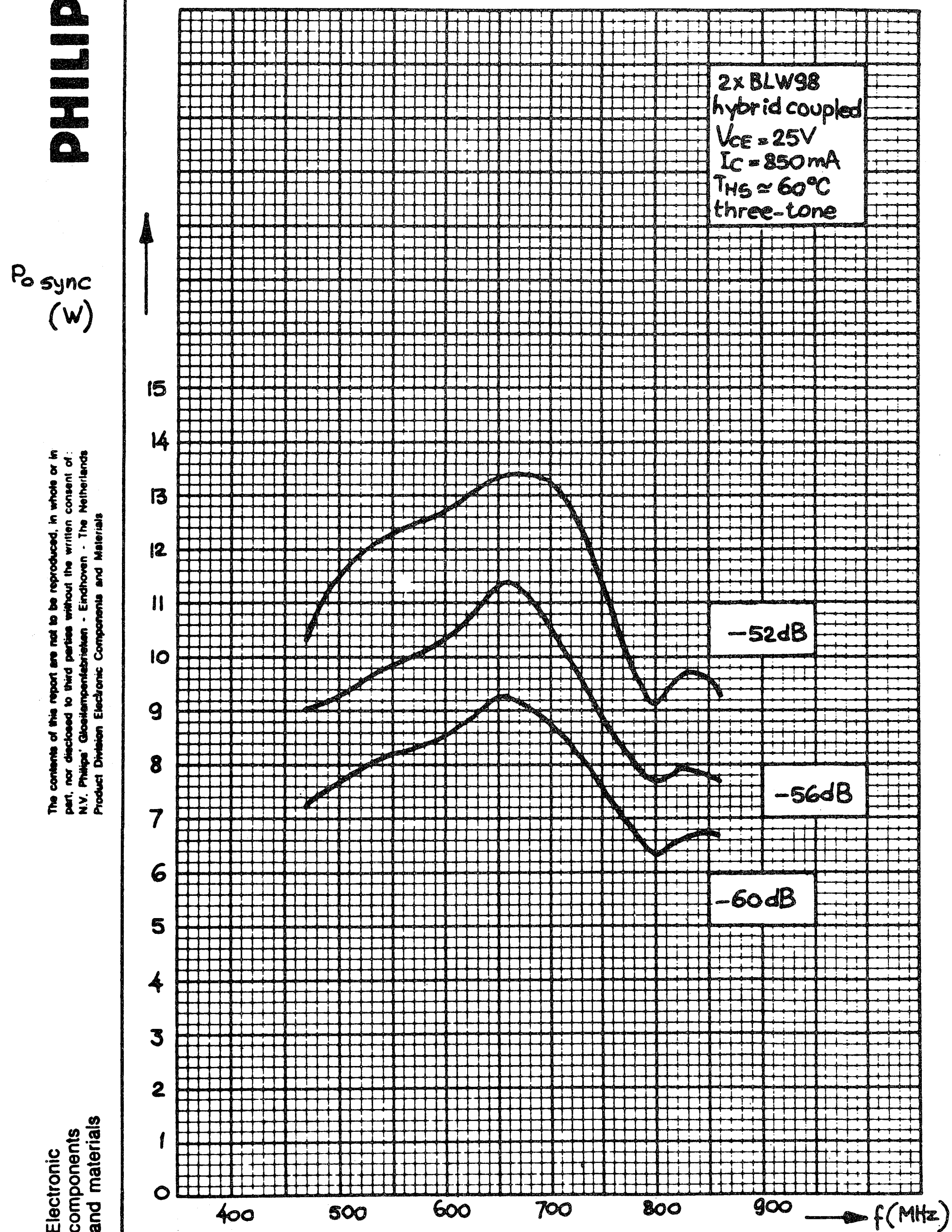

Fig.10

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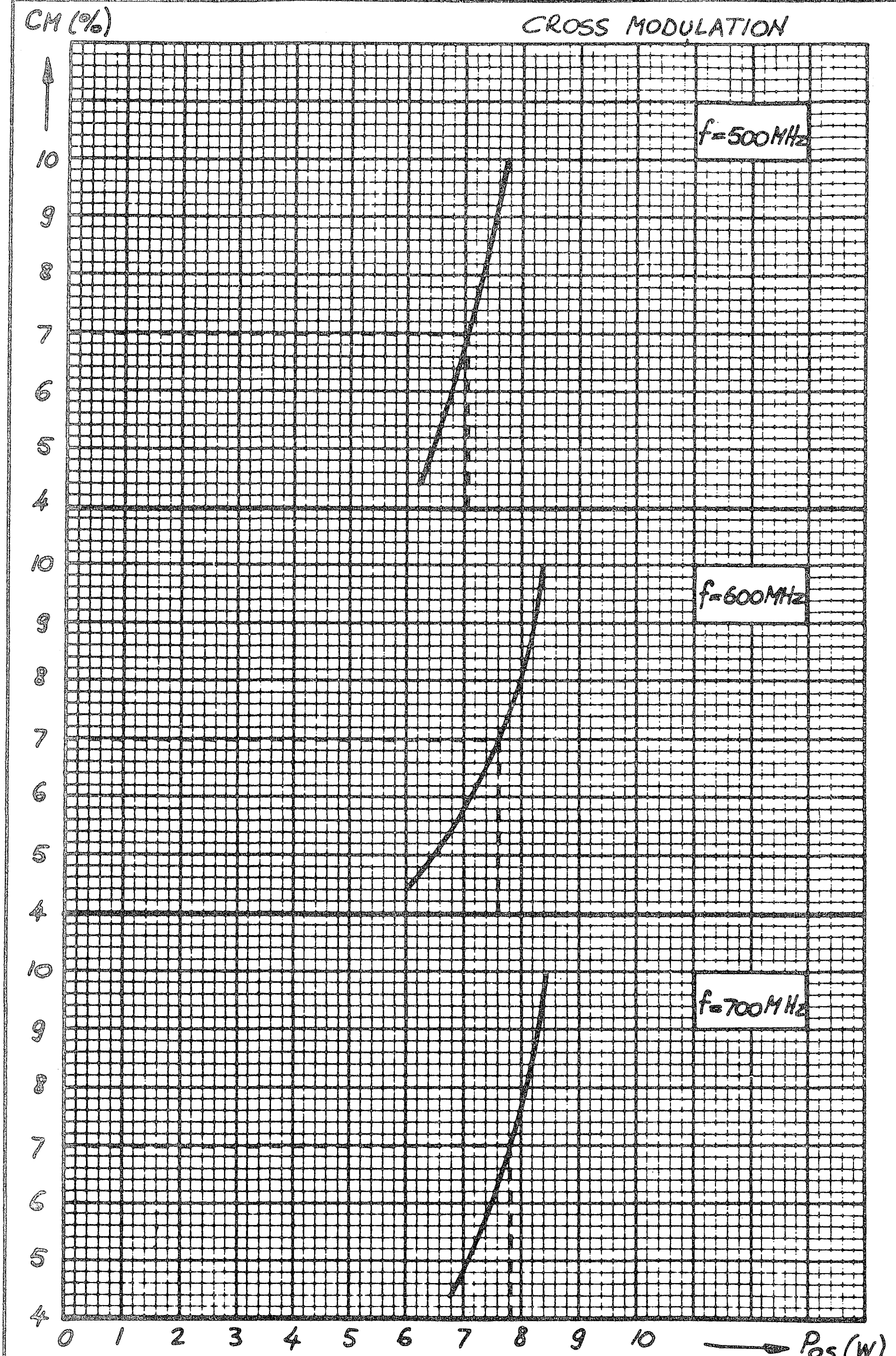

Fig.15.

Fig.16.

Fig.17.

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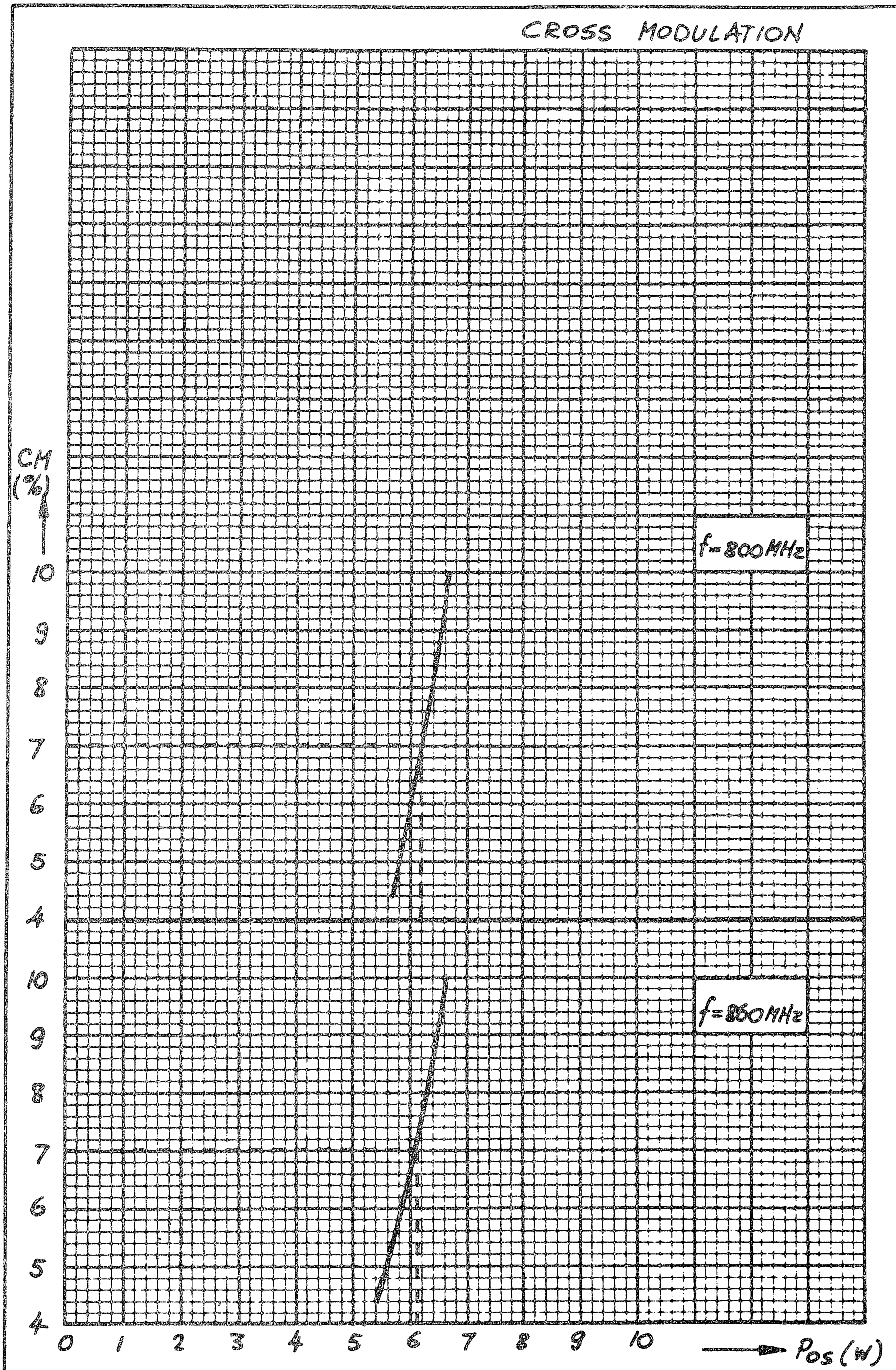

Fig.18.

Fig.19.

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laboratory report

number : EC08003 **date** : 21-08-1980

title : Wide-band linear power amplifier (174-230 MHZ) with 2 transistors BLV31.

author : A.Boekhoudt/R.F.F.Zwanen

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laboratory report

central application laboratory CAB
eindhoven - the netherlands

number : ECO 8003 date : 21-08-1980
project : 6967 pages: A1 ; R16 ;

title WIDEBAND LINEAR POWER AMPLIFIER (174-230 MHZ) WITH 2 TRANSISTORS BLV 31 (Redesign of ECO 7903)

author A. Boekhoudt, R.F.F. Zwanen

ABSTRACT

For application in driver or final stages of TV-transposers in band III (174-230 MHz) a linear wideband power amplifier has been designed with 2 transistors BLV31, coupled by means of 3dB-90^0 hybrids.
Each transistor is adjusted in class-A at V_{CE} = 22,5V and I_C = 0,9A.
The peak sync output power was 10W for a 3-tone I.M. distortion of -60dB. At this power level the cross-modulation was 8%.
The power gain was 16,2 $\pm$ 0,6dB.

Appr. R.A. Pölzl

Advies Octrooi d.d. 1980-08-28	XAV	GV		B		BL
Opgave Mamo d.d. 1980-08-25	XAV	XGV	XSP	B		BL
Datum: 25 AUG. 1980	Mamo					

1. INTRODUCTION

This report describes a linear wide-band amplifier for T.V. transposers in band III (174-230 MHz). The amplifier, containing 2 transistors BLV31, can deliver ε0L. eaP for -60dB intermodulation distortion (IMD) when operating in class A.
The power gain is appr. 16dB
The BLV31 is encapsulated in a ¼ inch capstan envelope with ceramic cap.

2. THEORETICAL CONSIDERATIONS

2.1. The equivalent circuit of the BLV31

In a first approach the transistor parameters such as gain, input- and load impedances are calculated at a class A adjustment of V_{CE} = 25V and I_c = 0,8A.
Table 1 gives the above mentioned transistor parameters for three frequencies. On these figures the amplifier has been designed.

f (MHz)	Gain (dB)	$R_i + j\ X_i$ (ohm)	$R_L + j\ X_L$ (ohm)
174	16,5	0,90 0,59	11,4 11,0
202	15,2	0,90 0,75	9,7 10,7
230	14,1	0,90 0,90	8,2 10,2

Table 1

A simplified equivalent circuit is shown in fig. 1.

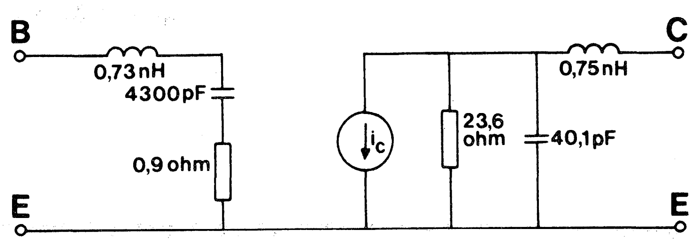

Figure 1

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2.2. The output network

The reactance of the collector choke being 130nH has been chosen about 7 times the parallel equivalent load resistance.
This reactance is so high that it has practically no influence on the h.f. performance.
The collector choke is connected at a distance of 5 mm from the transistor by a stripline. The equivalent circuit is shown in figure 1A. Lengths of the transmission lines in the figures 1A, 2and 3 are given with air as dielectric.

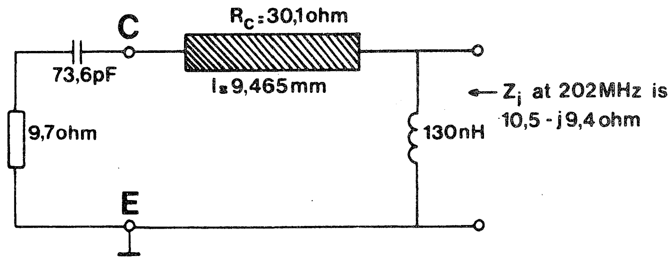
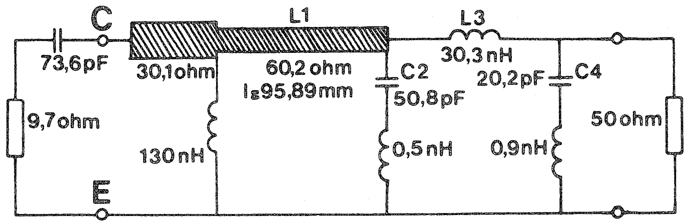


Figure 1A

The 10,5 ohm is matched to 50 ohm with a two section network which has been calculated according to Ref. 1.
The negative reactance of 9,4 ohm is tuned out by an equal positive reactance and this has to be added to L_1 in the calculation according to Ref. 1. The figures obtained by calculation are put in an optimization program which gives us the final output network; see figure 2.



$S_{max} \approx 1{,}16$

Figure 2

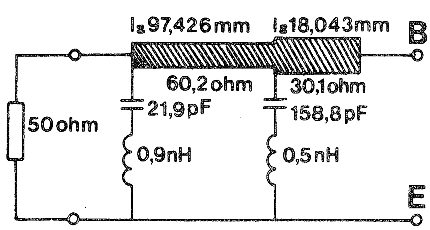
S_{max} is the maximum VSWR of the network calculated at the 50 ohms terminals.

2.3. The input network

The input network contains also two sections. Starting from the transistor, the first section has the function to make the overall gain constant. The gain slope of the transistor is 6dB per octave. So the first section has to have a positive-slope of 6dB per octave. The calculation of the first section has to be done on the top frequency of band III (230 MHz) and the loaded Q-factor must be approx. 4. This will result in an input impedance of $0{,}9 \times (4^2+1) = 15{,}3$ ohms.

The second section transforms the impedance from 15,3 ohm to 50 ohm;this calculation has to be done according to Ref. 1.

The final network after optimization is shown in figure 3, also given is the VSWR of the input at three frequencies.



VSWR	f(MHz)
5,2	174
2,6	202
1,6	230

Figure 3

3. THE PRACTICAL AMPLIFIER

The schematic diagram of the complete amplifier is shown in fig. 4. It contains 2 branches which are coupled by means of 3dB - 90^0 hybrids. An important property of these hybrids is that if the 2 output ports are loaded with equal impedances the input VSWR is always equal to 1.

The amplifier is designed on p.c. board with epoxy fibre-glass as a dielectric ($\varepsilon_r \approx 4{,}5$), the thickness of the board is 1/16 inch and it is copper clad on both sides. Fig. 6 shows the board and fig. 7 the lay-out of the components. To get a good contact between upper and lower side, rivets have been used at several places.

The amplifier has been aligned with a single tone swept frequency signal producing an output power between 15 and 20W. During this process it appeared that some of the components had to be changed. The results can be found in fig. 4. and the parts list.
It also became clear that an other DC adjustment, viz. V_{CE} = 22,5V and I_C = 0,9A gave a better result on I.M.D.
Each transistor is biased by a unit as shown in fig. 5.

4. Measurements

4.1 S-parameters.

Power gain and input reflection are depicted in fig. 8.
It appears that the power gain is appr. 1dB higher than the calculated one. A probable explanation is that the actual emitter lead inductance is somewhat smaller than the originally assumed value of 0,5 nH.
Fig. 9 shows the reverse isolation and output reflection.

4.2 Cross-modulation and I.M.D.

The upper part of fig. 10 presents the cross-modulation for 3 different power levels.
Similar information on the 3 tone I.M. distortion can be found in the lower part of this figure.
The measuring signals are defined in the figure.

5. REFERENCES

Ref. 1: G.L. Matthaei - Tables of Chebyshev Impedance - Transforming Networks of Low-Pass Filter Form. Proceedings of the IEEE August 1964, pp. 939-963.

Ref. 2: A.H. Hilbers and M.J. Köppen - A wide-band linear power amplifier (470-860 MHz) with two transistors BLW34. C.A.B. report ECO 7901 -

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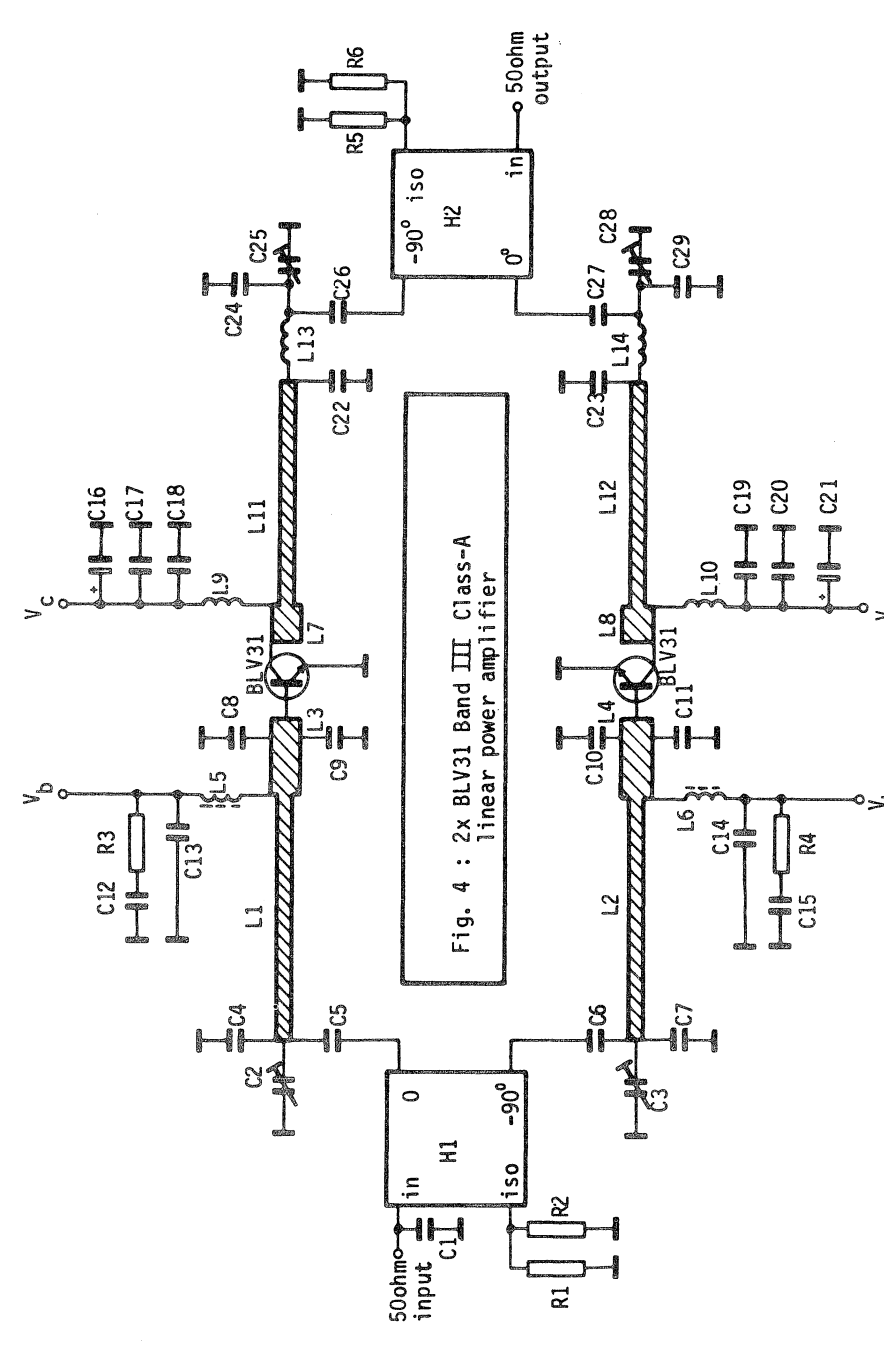

Fig. 4 : 2x BLV31 Band III Class-A linear power amplifier

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Parts list: BLV31 Band III Class A linear power amplifier (174 to 230 MHz)

C1 = 1,8 pF, chip capacitor
C2 = C3 = 1,8 to 10 pF, film dielectric trimmer (cat. no. 222280905002)
C4 = C7 = 18 pF, chip capacitor
C5 = C6 = 39 pF, chip capacitor
C8 = C9 = C10 = C11 = 100 pF, chip capacitor, connected at 5mm from transistor edge.
C12 = C15 = C17 = C20 = 330 nF, metalized film capacitor (cat. no. 222235225334)
C13 = C14 = C18 = C19 = 1000 pF, chip capacitor
C16 = C21 = 10 µF (40V), electrolytic capacitor (cat. no. 222212117109)
C22 = C23 = 22 pF, chip capacitor
C24 = C29 = 6,8 pF, chip capacitor
C25 = C28 = 1 to 3,5 pF, film dielectric trimmer (cat. no. 22228090501)
C26 = C27 = 100 pF, chip capacitor
(chip capacitors: ATC type 100B - C - MSX - 500)

R1 = R2 = R5 = R6 = 100 Ohm, power metal film resistor PR52 type (cat. no. 232219231001)
R3 = R4 = 10 Ohm, carbon resistor CR68 type

H1 = H2 = 3dB - 90^{0} coupler, model no. 10262 - 3, freq. range 125 to 250 MHz, ANAREN MICROWAVE INC.

L1 = L2 = 60 Ohm stripline, w = 2mm, 1 = 54,1mm
L3 = L4 = 30 Ohm stripline, w = 6mm, 1 = 9,5mm
L5 = L6 = 1 µH, microchoke
L7 = L8 = 30 Ohm stripline, w = 6mm, 1 = 5mm
L9 = L10 = 25nH, 2 turns enamelled Cu wire (1mm), int. diam. 5mm, length = 5mm, leads 2 x 3 mm
L11 = L12 = 60 Ohm stripline, w = 2mm, 1 = 53,2mm

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Parts list: BLV31 Class A linear power amplifier (174 to 230 MHz)

L13 = L14 = 38 nH, 4 turns enamelled Cu wire (1mm),
int. diam. 3,5mm, length 9mm, leads 2 x 3 mm

The striplines are printed on double Cu-clad printed circuit board with epoxy fibre-glass dielectric (ε_r = 4,5), thickness 1/16 inch.

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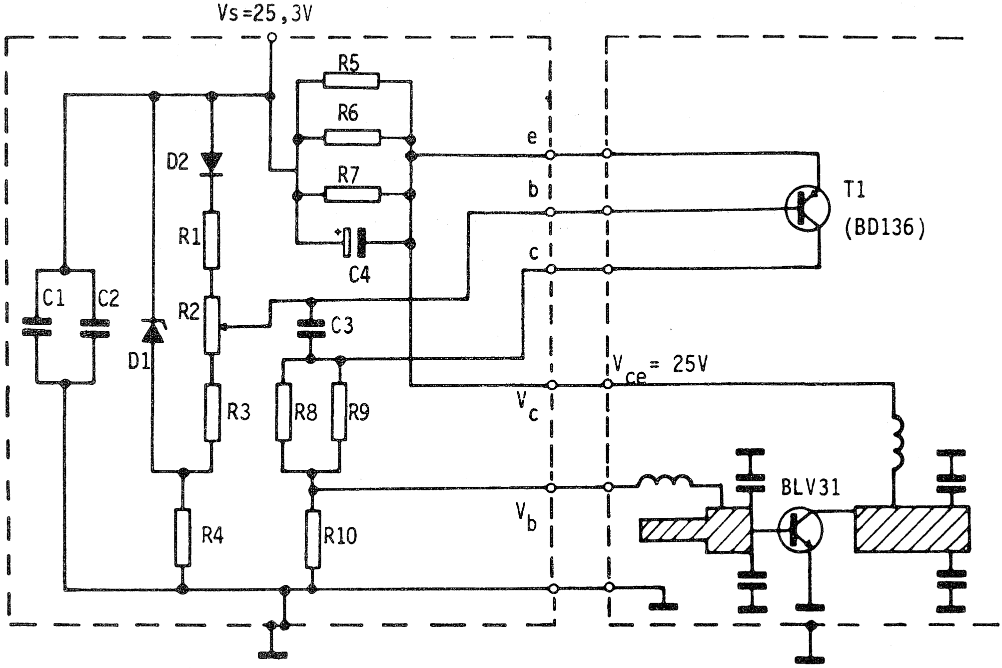

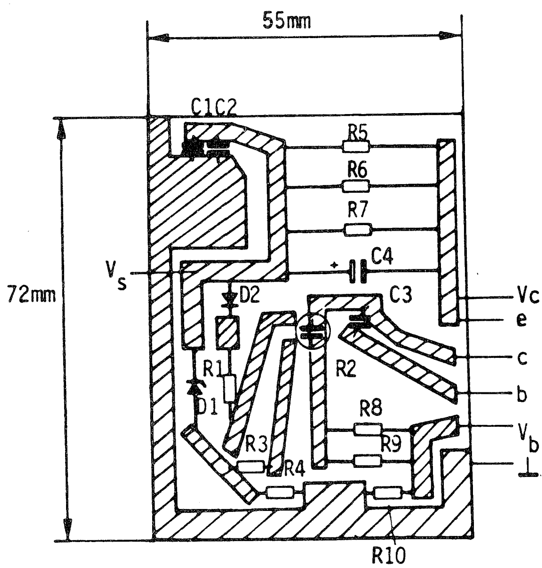

Fig. 5 : Class A bias circuit for a single transistor BLV31

Parts list: Class A bias circuit for a single transistor BLV31

R1 = 150 Ohm, carbon resistor CR25 type
R2 = 100 Ohm, preset potentiometer CTP10 type
R3 = 10 Ohm, carbon resistor CR25 type
R4 = 1000 Ohm, carbon resistor CR25 type
R5 = R6 = R7 = 8,2 Ohm, rectangular wirewound resistor EH 07 type
R8 = R9 = 180 Ohm, carbon resistor CR25 type
R10= 33 Ohm, carbon resistor CR25 type

C1 = C3 = 100nF, metalized film capacitor
C2 = 100 pF, ceramic capacitor
C4 = 10 µF, electrolytic capacitor

D1 = BZY 88 (3V3)
D2 = BY206
T1 = BD136

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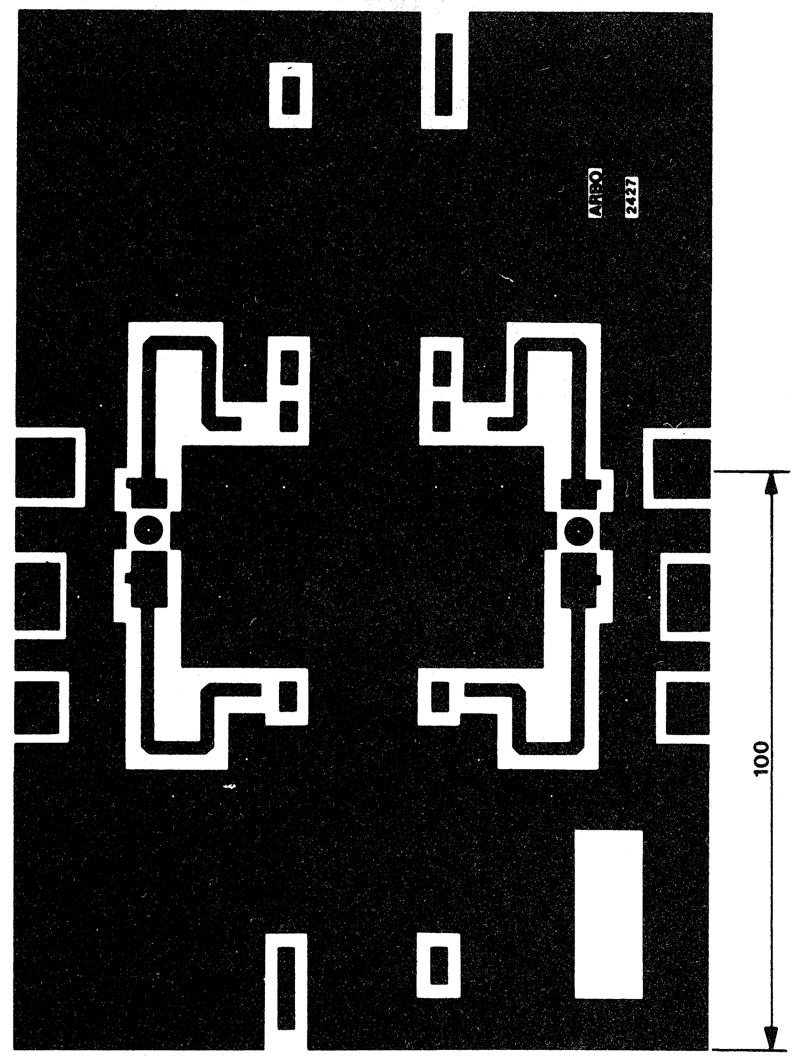

Figure 6

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Unfortunately the numbers in the lay-out of fig. 7 do not correspond with those of the schematic diagram of fig. 4. The reader is referred to the translation table below.

Number in lay-out (Fig. 7)	Number in schematic diagram (fig. 4)
R1	R3
R2	R5
R3	R6
C3	C2
C5	C4
C6	C5
C8	C8
C9	C9
C10	C12
C11	C13
C13	C22
C14	C16
C15	C17
C16	C18
C17	C24
C18	C25
C19	C26
L3	L5
L6	L9
L9	L13

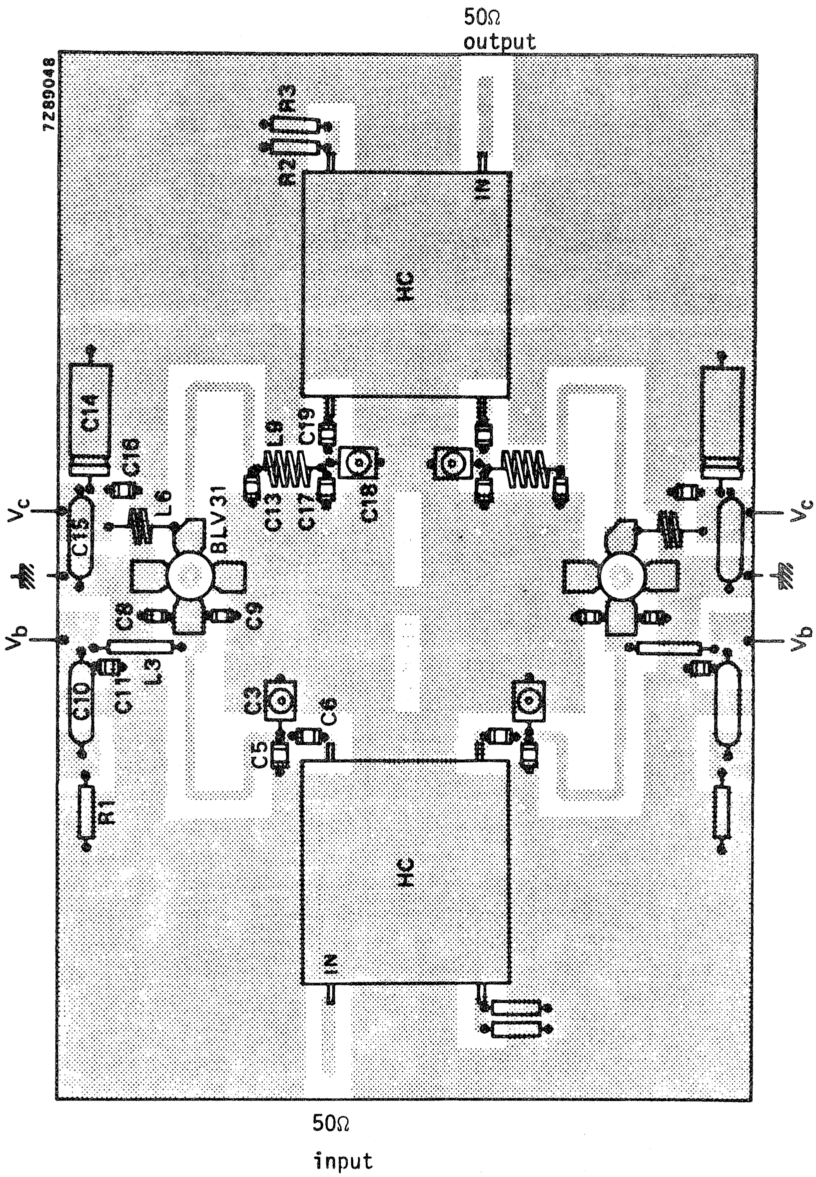

Fig. 7.

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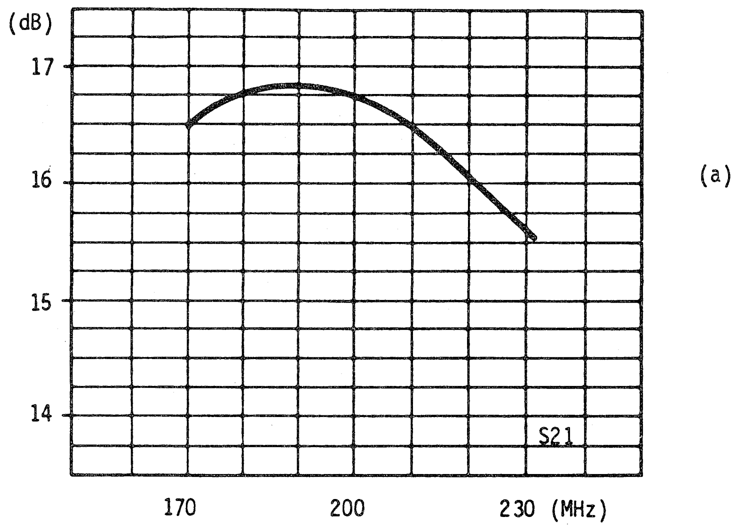


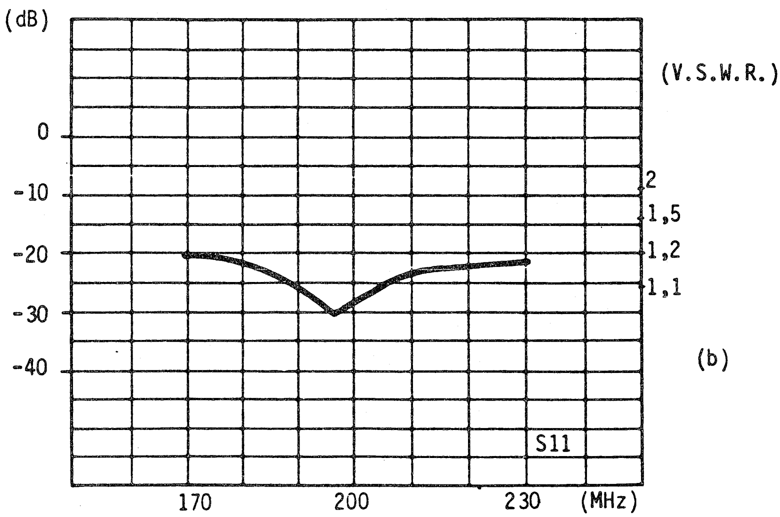
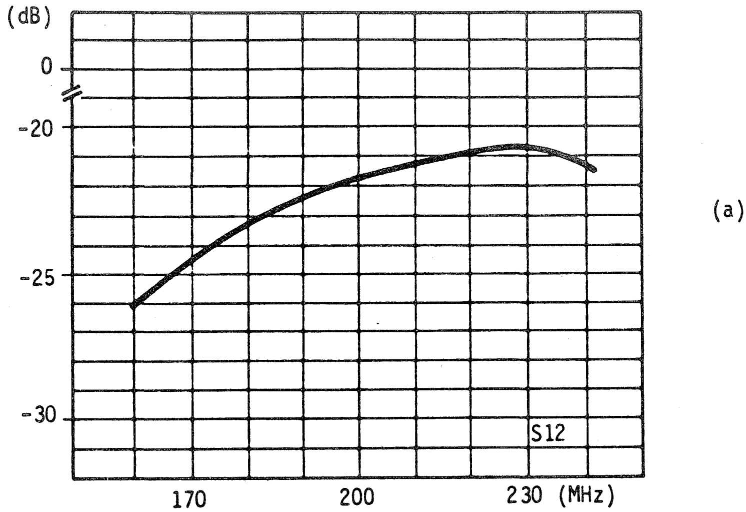


Fig. 8 : 2x BLV31 wideband Band III power amplifier

a) Forward transducer gain

b) Input voltage standing wave ratio



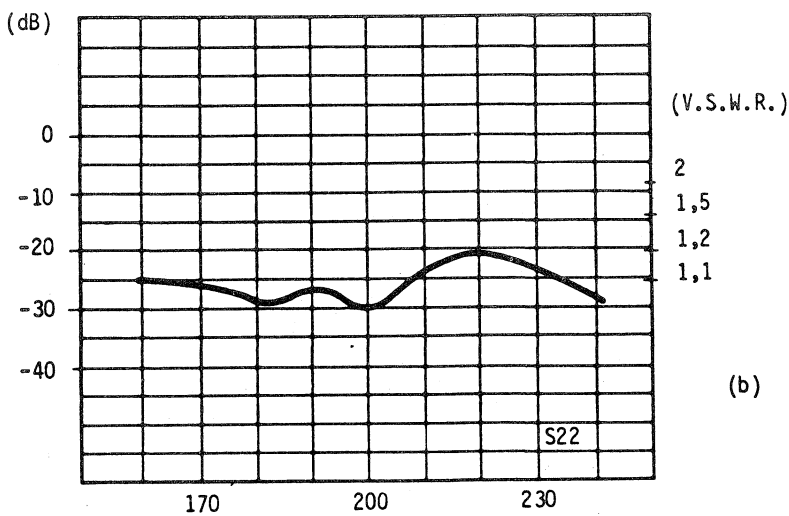
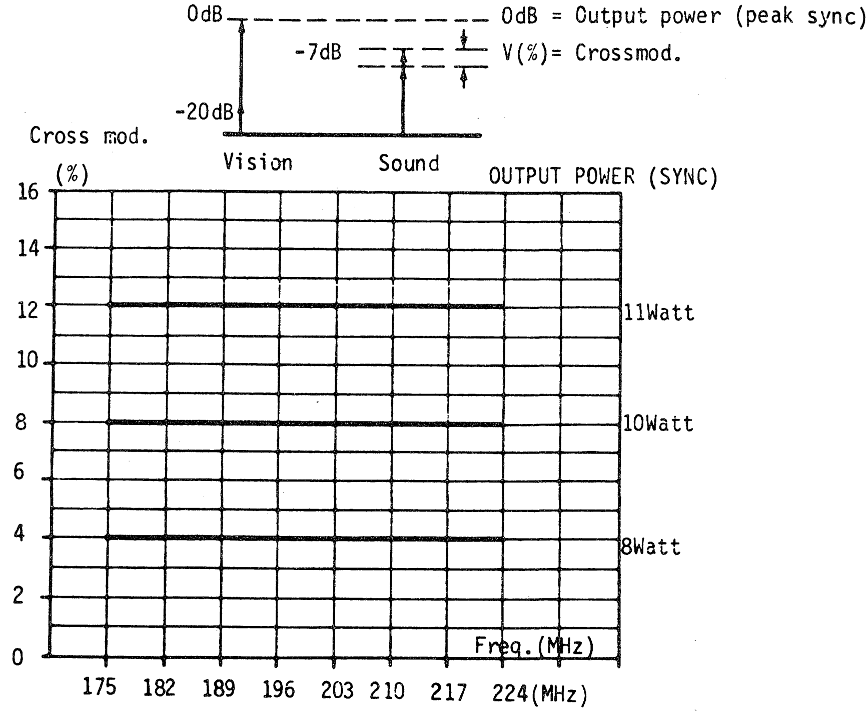


Fig. 9 : 2x BLV31 wideband Band III power amplifier

a) Reverse transducer gain

b) Output voltage standing wave ratio

2x BLV31 : V_{CE} 22,5V : I_C = 2x 0,9A



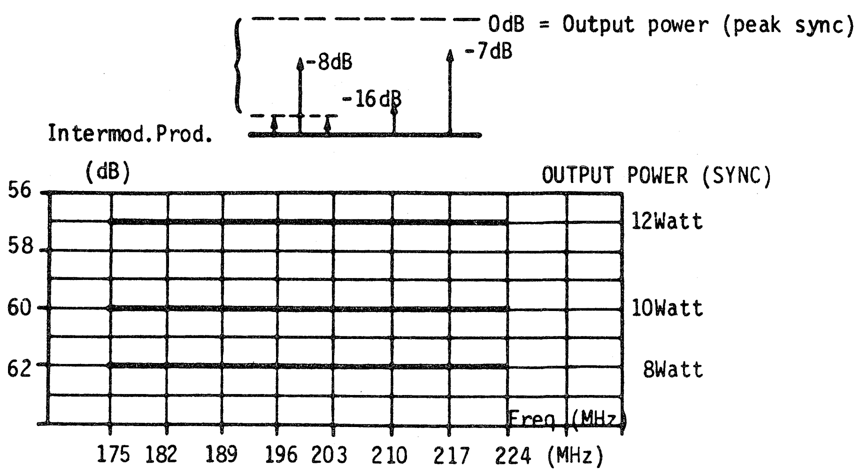


Fig. 10 Crossmod. and intermod. prod. of the 2x BLV31 wideband Band III linear power amplifier.

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laboratory report

number : EC08005

date : 27-08-1980

title : A WIDE BAND CLASS A LINEAR POWER AMPLIFIER (170-230MHz) WITH 2 TRANSISTORS BLV33F.

author : R.F.F.Zwanen

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laboratory report

central application laboratory CAB
eindhoven - the netherlands

number: ECO 8005 date: 27-08-1980

project: 1068 pages: A1 : R16 :

title A WIDE-BAND CLASS-A LINEAR POWER AMPLIFIER (170-230MHz) WITH 2 TRANSISTORS BLV33F

author R.F.F. Zwanen (Dev. Transm. and Microw. dev.)

ABSTRACT

For application in driver or final stages of TV-transposers in Band III (174-230MHz) a linear wideband power amplifier has been designed with 2 transistors BLV33F, coupled by means of 3dB-90^{o} hybrids.
Each transistor is adjusted in Class-A at V_{CE} = 25V and I_C = 3,25A.
A demonstration model showed a peak sync. output power of 40W at a 3-tone I.M. distortion between -52 and -53dB.
At this power level the cross modulation varied from 15 to 18%.
The power gain is between 13,3 and 13,6 dB.
For natural convection cooling the heatsink temperature is 40^{o}C above ambient temperature.

Appr. R.A. Pölzl

Advies Octrooi d.d. 1980-09-08	XAV	GV		B		BL	
Opgave Mamo d.d. 1980-08-28	XAV	XGV	XSP	B		BL	
Datum: 28 AUG. 1980	Mamo						

1. INTRODUCTION

For application in T.V. transposers and transmitters for Band III a wideband linear power amplifier has been designed with 2 transistors BLV33F, coupled by means of 3dB -90^0 hybrids. Each transistor is adjusted in Class-A at V_{CE} = 25V and I_C = 3,25A.

Note: The BLV33F is a high gain, internally matched, 1/2" 6 leads flange version of the BLV33.

2. DESIGN OF THE AMPLIFIER

For class-A operation the BLV33F is specified at V_{CE} = 25V, I_C = 3,25A.
The corresponding typical gain, input and load impedance are given below:

Freq. (MHz)	Gain (dB)	Input impedance (Ohm)	Load impedance (Ohm)
174	12,94	0,85 + j0,59	2,68 + j1,24
202	12,88	1,02 + j0,47	2,23 + j0,90
230	13,91	0,93 + j0,02	1,84 + j0,51

To obtain a high linear output and at the same time good input and output matching (V.S.W.R. $\leq$ 1,2) 3dB - 90^0 hybrids are used. The reflected input power will be absorbed in the 50 Ohm resistor, matching the isolated port (see fig. 1).
For detailed information on computer-aided design (carried out by Mr. Hilbers Central Application Laboratory) see ref. 1-2-3-4.
The transistors used in this particular amplifier are typical products, measured in a narrow band test amplifier and specified as follows:

$V_{CE} = 25V - I_C = 3.25A - Th = 70^0$

Transistor type	: BLV33F		
Batch no.	: MD 8-16 no.7; MD 8-16 no. 10		
Vision frequency	:	224,25 MHz	
Output power (peak sync)	:	17,7W	18W
Intermod. product	:	-55dB	-55dB
Gain	:	14,2dB	14,2dB

3. ADJUSTMENTS OF THE AMPLIFIER

The amplifier consists of two equal BLV33F branches (see fig. 1) and both transistors are separately biased at $V_{CE} = 25V - I_C = 3,25A$.

The printed circuit board of the 2 x BLV33F wideband amplifier is given in fig. 2 and schematic diagram + lay-out of the bias unit is given in fig. 3.*)
Each branch was adjusted for maximum and flat gain by means of a high power sweep with a frequency range from 170 to 230MHz.
The output of the amplifier was leveled at 40W which means about 50% of the D.C. input power.
After that, both branches are coupled by means of $3dB-90^0$ hybrids.

4. ASSEMBLING OF THE AMPLIFIER AND MECHANICAL DATA

Due to the dimensions of the printed circuit board (220 x 210mm) 2 extruded blackened aluminium heat sinks (cat. no. 56293) are screwed on an aluminium plate (thickness 12mm) which on its turn is screwed on the heat sink.
Special attention has been paid to the surface finishing to keep the thermal resistance as low as possible.
Dimensions of the amplifier:
l = 224mm - w = 223mm - h = 113mm. Weight: 7,5 kg.

*) Fig. 9 at the end of the report shows the lay-out of the amplifier with the situation of the components.

5. MEASURED RESULTS

In fig. 4 the typical results of crossmodulation and 3-tone intermod. product (from 170 to 230MHz) have been given for peak sync output powers of 30W and 40W.
Fig. 5 shows peak sync output power as function of 3-tone intermod. products (measured on channel 12: Vision freq. 224,25 MHz - Sound freq. 229,75MHz).
In fig. 6-7 the forward and reverse transducer gain as well as input and output voltage standing wave ratio are given.
The measuring test set-up is depicted in fig. 8.
Note: Signal levels 3-tone measurements:
Vision carrier -8dB; Sound carrier -7dB;
Sideband -16dB; 0dB corresponds to peak sync.

Signal levels crossmodulation:
Vision carrier switched from -20dB to 0dB;
Sound carrier -7dB; 0dB = peak sync level
Crossmodulation is defined as the voltage variation (%) of the sound carrier.

6. CONCLUSION

Two transistors BLV33F, coupled by means of 3dB -90^o hybrids, can deliver an output power (peak sync) of typ. 40W for -52dB three-tone intermodulation.
At 40W output the crossmodulation varied from 15 to 18% in Band III (170 to 230 MHz)
The gain of the amplifier is typically 13,3dB + 0,3dB.
The required D.C. input is approx. 165W.
Using a high power sweep with adjustable transistor output leveling provides a suitable methode to adjust a linear wideband power amplifier.

7. REFERENCES

Ref. 1: G.L. Matthaei - Tables of Chebyshev
Impedance Transforming Networks of Low-Pass Filter Form.
Proceedings of the IEEE August 1964
pp 939 - 963.

Ref. 2: A.H. Hilbers and M.J. Köppen - A wideband linear power amplifier (470 - 860 MHz) with two transistors BLW34.
C.A.B. report ECO 7901.

Ref. 3: A. Boekhoudt - Wideband linear power amplifier with two transistors BLV31.
C.A.B. report ECO 8003.

Ref. 4: R.F.F. Zwanen - A wideband Class-A linear power amplifier (170 - 230 MHz) with two transistors BLV33. C.A.B. report ECO 7904.

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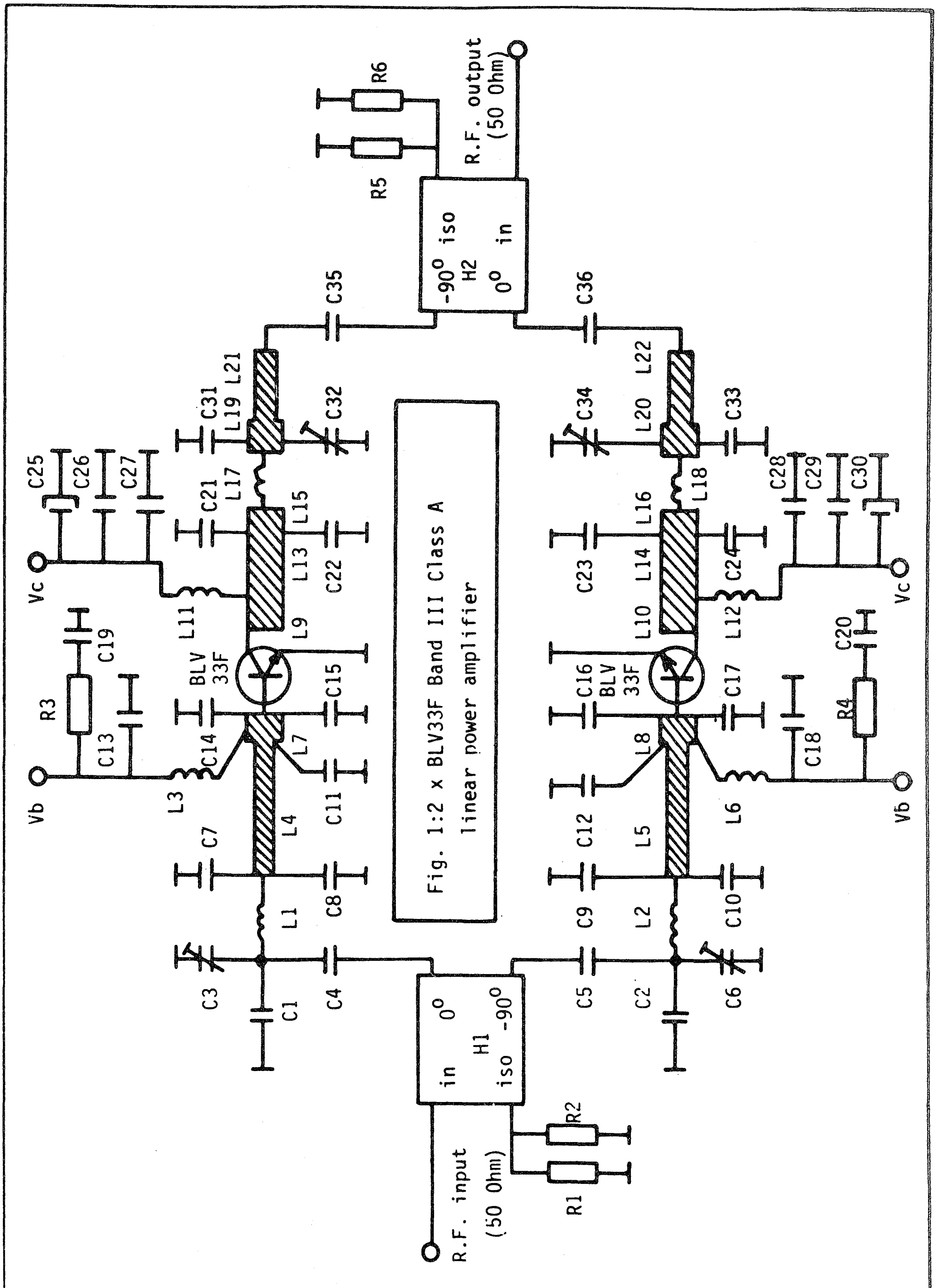

Fig. 1:2 x BLV33F Band III Class A linear power amplifier

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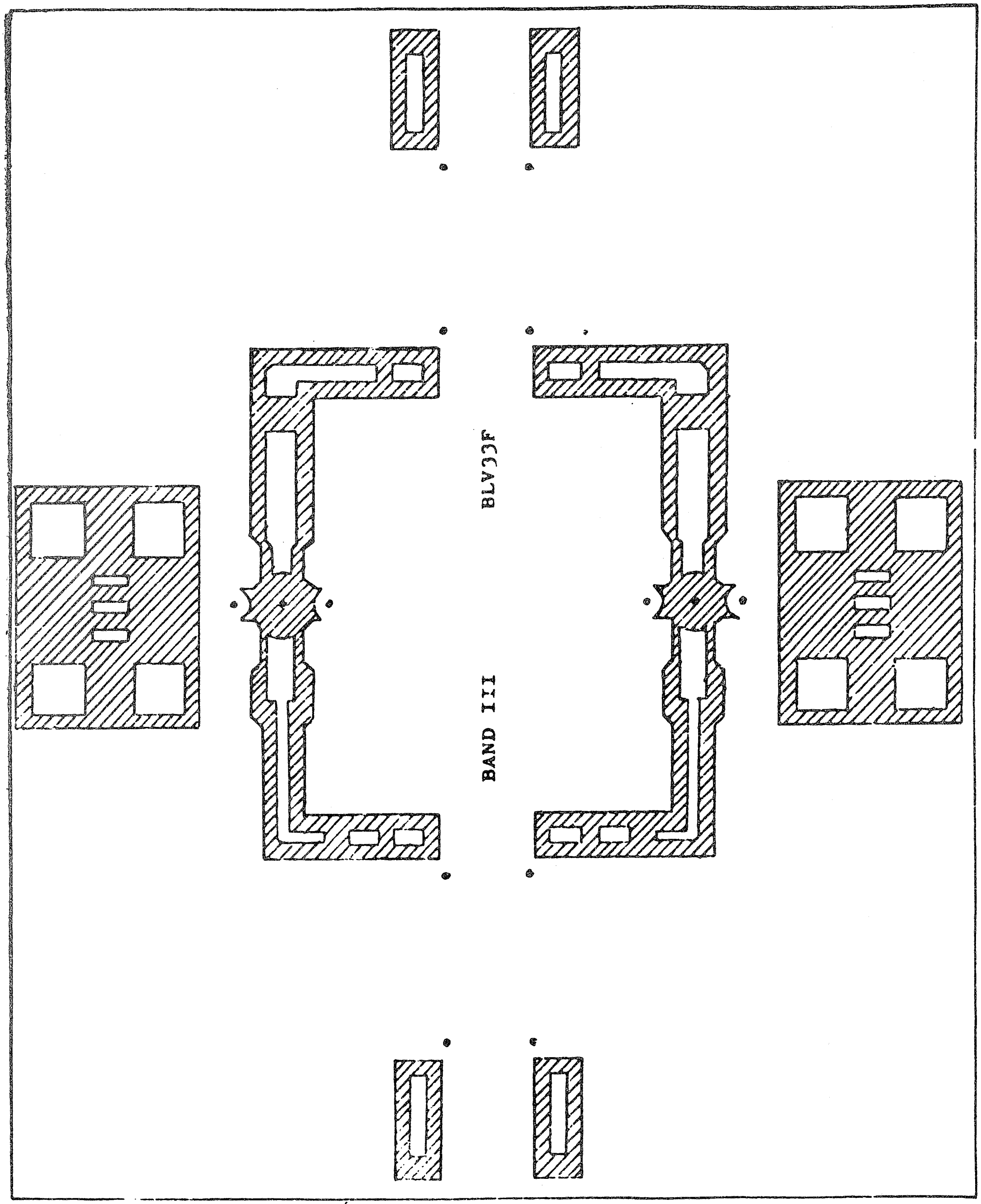

Fig. 2: Printed circuit board 2 x BLV33F wideband power amplifier.

Parts list: BLV33F Band III Class A linear power amplifier (170 to 230 MHz)

C1 = C2 = C7 = C9 = 10 pF, chip capacitor
C3 = C6 = C32 = C34 = 1,8 to 10 pF, film dielectric trimmer (cat. no. 222280905002)
C4 = C5 = C35 = C36 = 220 pF, chip capacitor
C8 = C10 = 39 pF, chip capacitor
C11 = C12 = 68 pF, chip capacitor
C13 = C18 = C27 = C28 = 1000 pF, chip capacitor
C14 = C15 = C16 = C17 = 120 pF, chip capacitor
C19 = C20 = C26 = C29 = 300nF, metalized film capacitor (cat. no. 222235225334)
C21 = C22 = C23 = C24 = 56pF, chip capacitor
C25 = C30 = 10 uF (40V), electrolytic capacitor (cat. no. 222212117109)
C31 = C33 = 18pF, chip capacitor
(chip capacitors: ATC type 100B - C - MSX - 500)

R1 = R2 = R5 = R6 = 100 Ohm, power metal film resistor PR52 type (cat. no. 232219231001)
R3 = R4 = 10 Ohm, carbon resistor CR68 type

H1 = H2 = 3dB - 90^{0} coupler, model no. 10262 - 3, range 125 to 250 MHz, ANAREN MICROWAVE INC.

L1 = L2 = 25 nH; 2 turns enamelled Cu wire (1mm); int. diam. 5 mm; leads 2 x 3 mm.
L3 = L6 = 90 nH; 5 turns closely wound enamelled Cu wire (1,5mm) int. diam. 6,5 mm; length 5 mm; leads 2 x 9 mm.
L4 = L5 = 60 Ohm stripline; w = 2 mm; length = 30 mm.
L7 = L8 = 30 Ohm stripline; w = 6 mm; length = 11 mm.
L9 = L10 = 40 Ohm stripline; w = 4 mm; length = 5 mm.
L11 = L12 = 20 nH; Cu strip (1mm); length = 17 mm; h = 5 mm;w=4mm.
L13 = L14 = 30 Ohm stripline; w = 6 mm; length = 17 mm.
L15 = L16 = 30 Ohm stripline; w = 6 mm; length = 4mm.

L17 = L18 = 28 nH; 2 turns enamelled Cu wire (1,5mm); int. diam. 6,5 mm; length 9 mm; leads 2 x 3 mm.
L19 = L20 = 30 Ohm stripline; w = 6 mm ; length = 6mm.
L21 = L22 = 50 Ohm stripline; w = 3 mm; length = 15 mm.

The stiplines are printed on double Cu-clad printed circuit board with epoxy fibre-glass dielectric (ε_r = 4,5); thickness 1/16 inch.

Parts list: Class A bias circuit for a single transistor BLV33F

R1 = 150 Ohm, carbon resistor CR25 type
R2 = 100 Ohm, preset potentiometer CTP10 type
R3 = 10 Ohm, carbon resistor CR25 type
R4 = 1000 Ohm, carbon resistor CR25 type
R5 = R6 = R7 = 1,8 Ohm, rectangular wirewound resistor EH707 type
R8 = R9 = 180 Ohm, carbon resistor CR25 type
R10 = 33 Ohm, carbon resistor CR25 type

C1 = C3 = 100 nF, metalized film capacitor
C2 = 100pF, ceramic capacitor
C4 = 10 uF, electrolytic capacitor

D1 = BZY 88 (3V3)
D2 = BY 206
T1 = BD 136

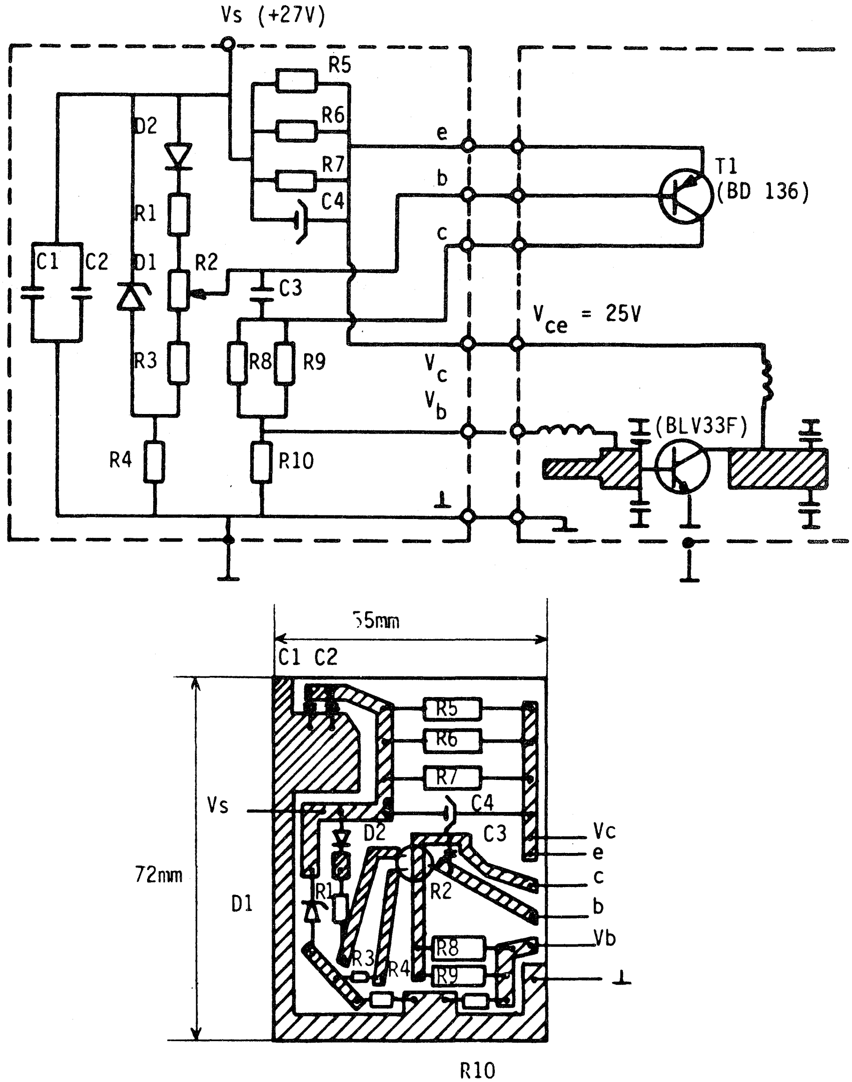

Fig. 3: Class A bias circuit for a single transistor BLV33F

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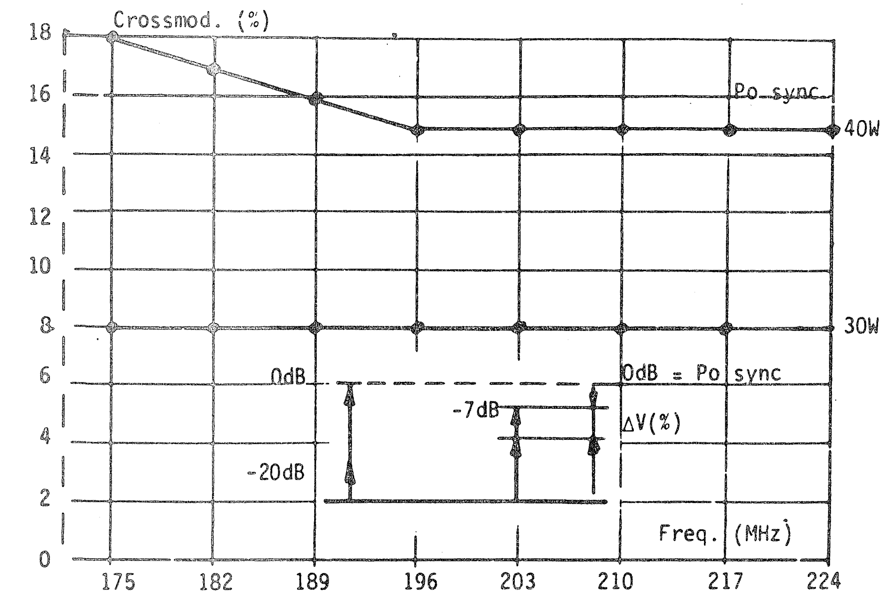

2 x BLV33F Class A linear power amplifier

Voc -25V - Ic = 2 x 3,25A - Tamb - 23°C - Th = 65°C

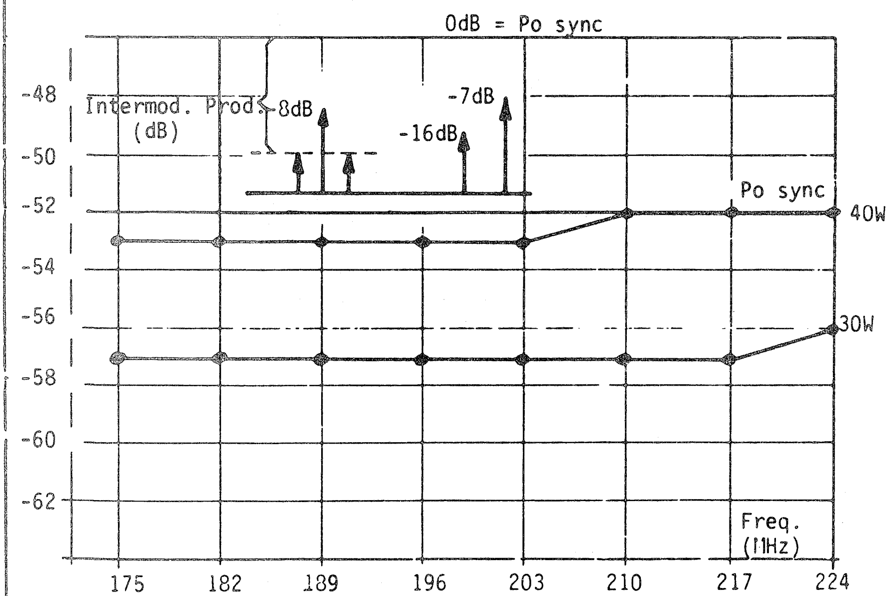

Fig. 4: Crossmod. and intermod. prod. of the 2 x BLV33F wideband Band III linear power amplifier.

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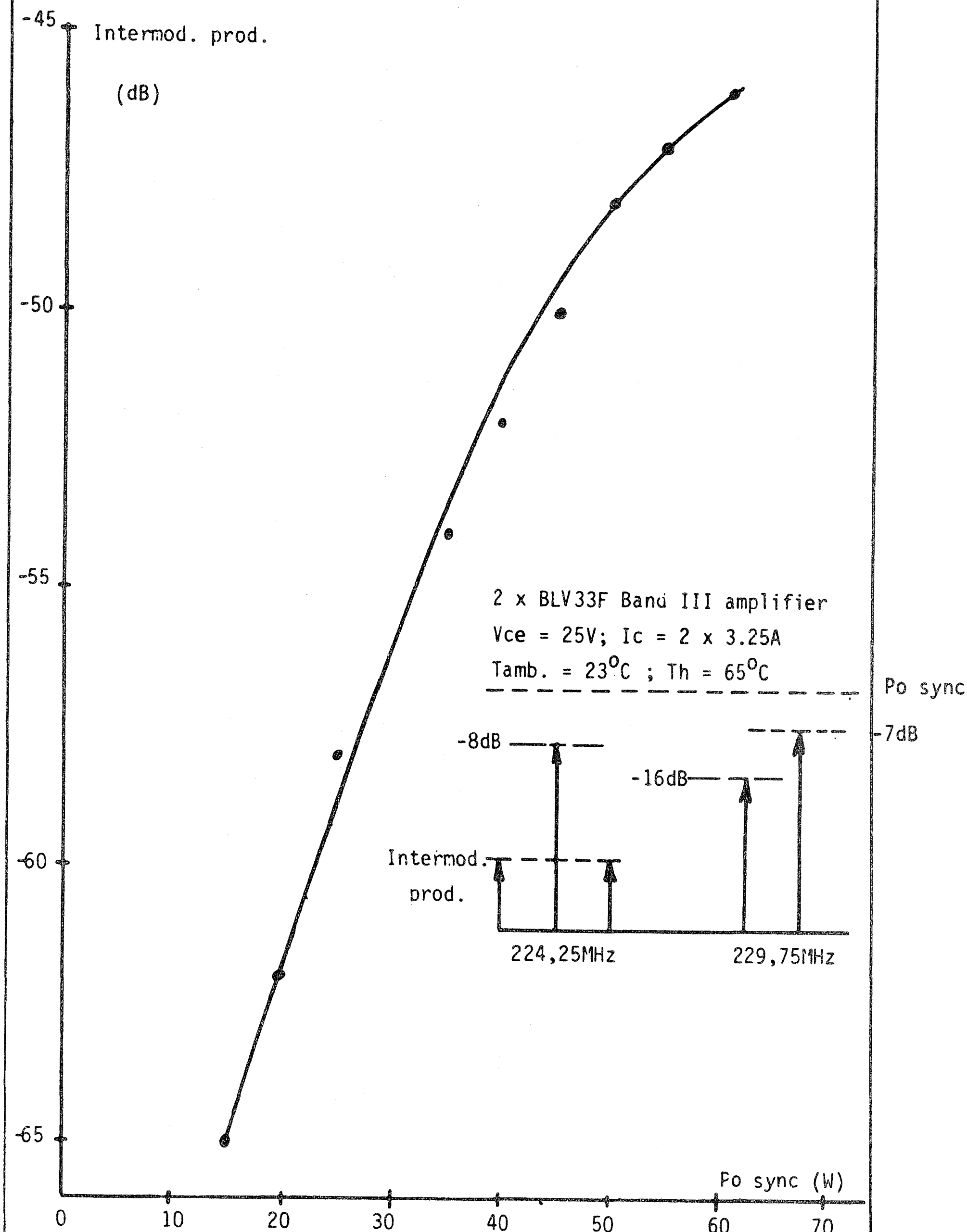

Fig. 5: Three-tone intermod. product as function of Po sync
Vision freq. 224 MHz; sound freq. 229,5 MHz.

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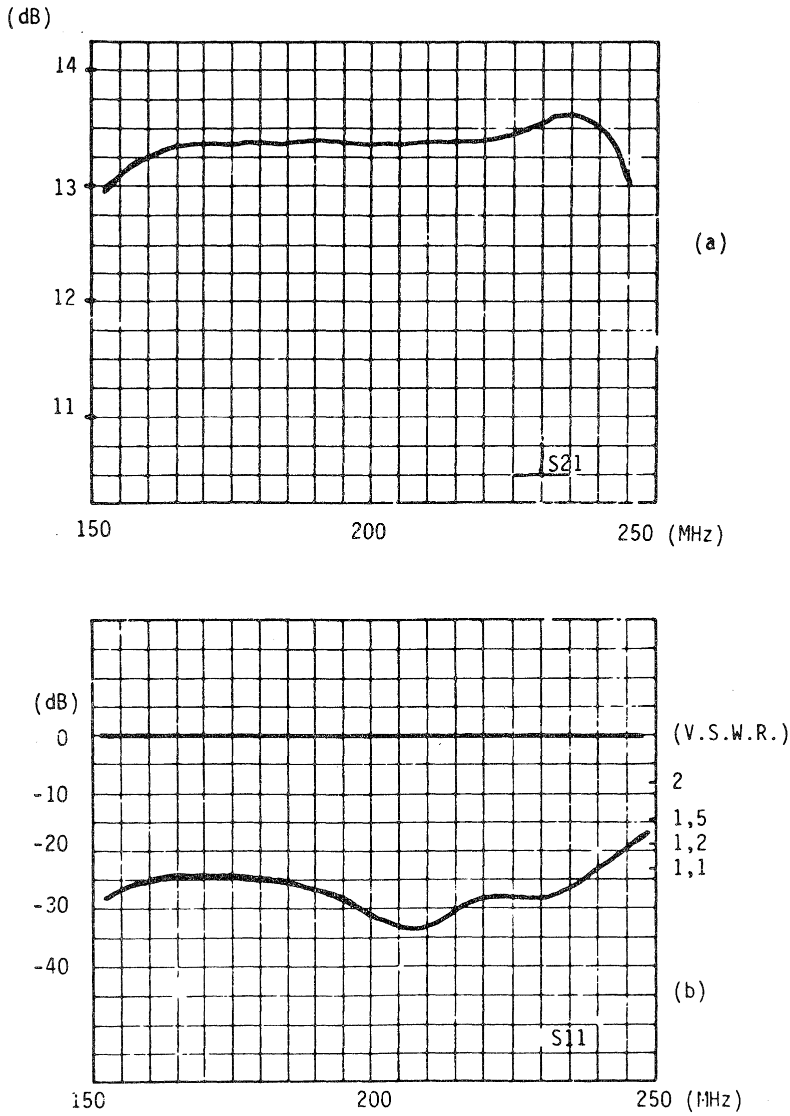

Fig. 6. 2 x BLV33F wideband Band III power amplifier
a) Forward transducer gain
b) Input voltage standing wave ratio.

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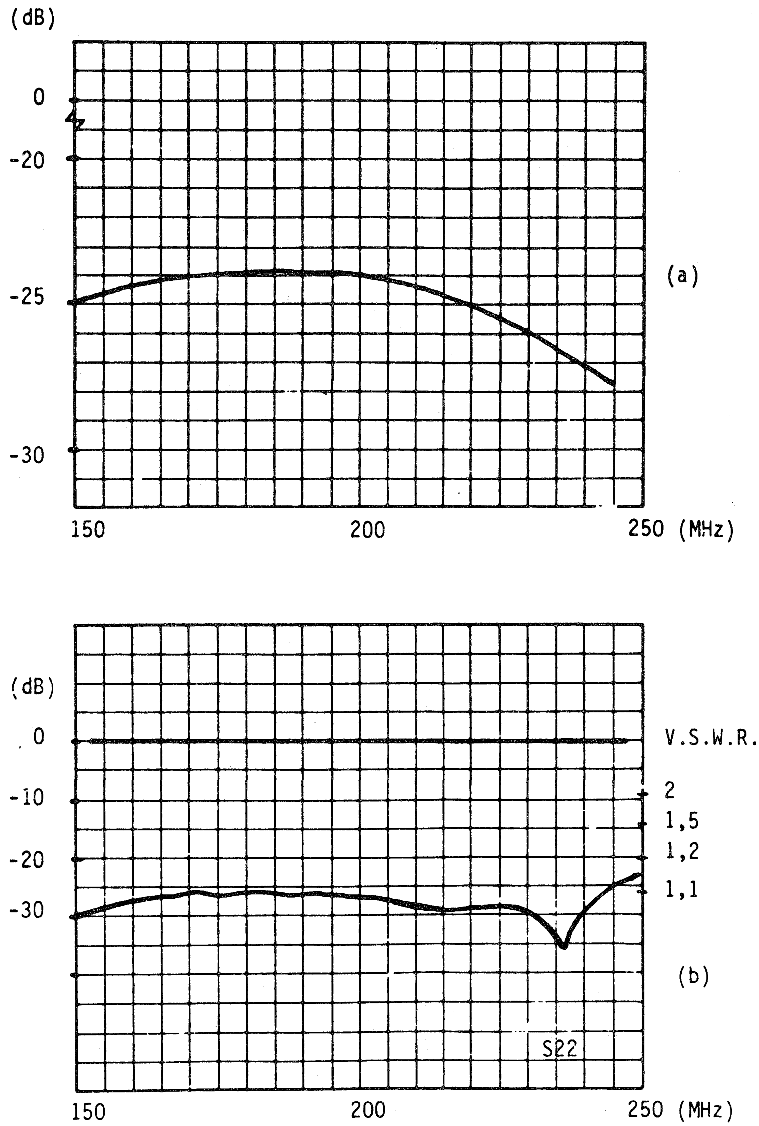

Fig. 7: 2 x BLV33F wideband power amplifier
a) Reverse transducer gain
b) Output voltage standing wave ratio

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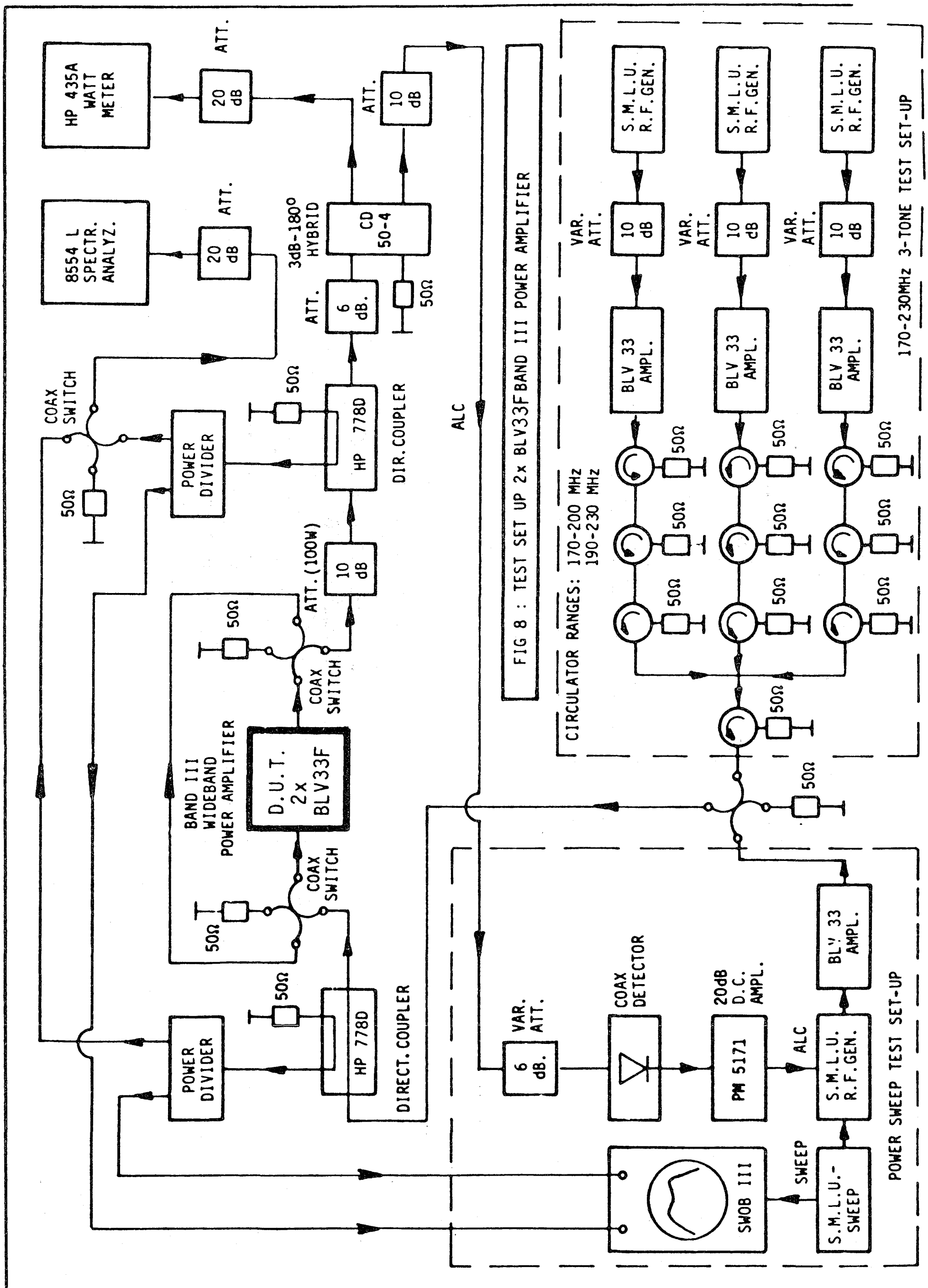

FIG 8 : TEST SET UP 2x BLV33F BAND III POWER AMPLIFIER

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Unfortunately the numbers in the lay-out of fig. 9 do not correspond with those of the schematic diagram of fig. 1. The reader is referred to the translation table below.

Number in lay-out (fig. 9)	Number in schematic diagram (fig. 1)
R1	R3
R2	R5
R3	R6
C2	C1
C3	C3
C4	C4
C5	C7
C6	C8
C7	C11
C8	C14
C9	C15
C10	C19
C11	C13
C12	C21
C13	C22
C14	C25
C15	C26
C16	C27
C17	C31
C18	C32
C19	C35
L1	L1
L3	L3
L6	L11
L9	L17

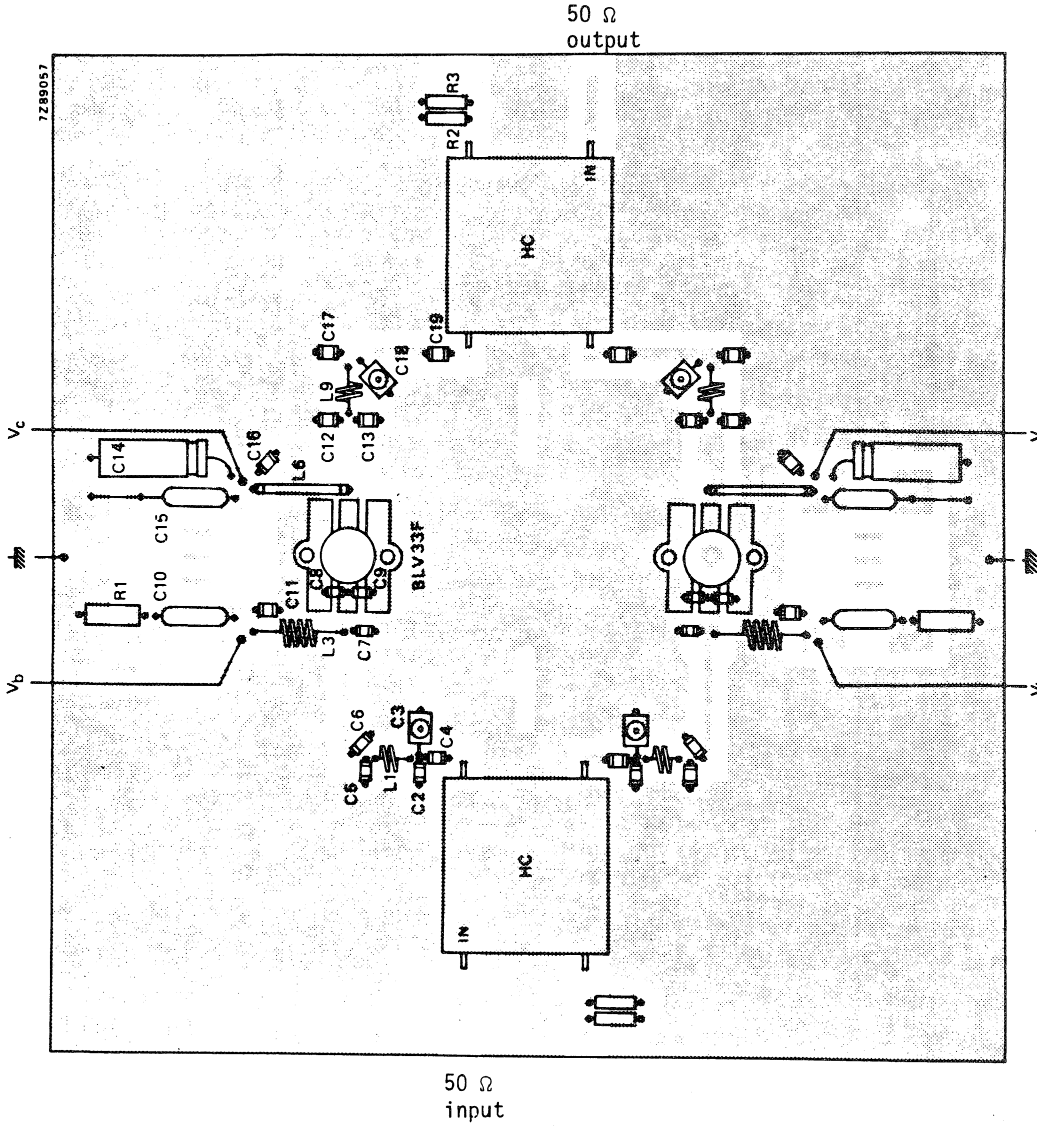

Fig. 9: Lay-out of amplifier with situation of components.

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laboratory report

number : EC08006 date : 24-09-1980

title :

DESIGN OF LINEAR POWER AMPLIFIERS FOR TV BANDS III IV V WITH THE BFQ68.

author : A.H.Hilbers.

Electronic components and materials

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laboratory report

central application laboratory CAB
eindhoven - the netherlands

number: ECO 8006 date: 24-09-1980

project: 6965 pages: A1 ; R11 ;

title DESIGN OF LINEAR POWER AMPLIFIERS FOR TV BANDS III-IV-V WITH THE BFQ68.

author A.H. Hilbers

ABSTRACT

Theoretical designs of wideband amplifiers for TV transposer service in bands III-IV-V are given.
The amplifiers use the BFQ68 adjusted in class-A at V_{CE} = 15V and I_C = 240mA.
Three designs have been made. The table below shows the expected performance for the case that 2 amplifier branches are combined*).

Freq. range (MHz)	470 - 860	174 - 230	174 - 860
Power gain (dB)	10,1 $\pm$ 0,1	18,4 $\pm$ 0,1	9,7 $\pm$ 0,2
P_o (W_{ps}) at IMD = -60 dB	1,38	1,48	1,29

Appr. R.A. Pölzl

*) By means of 3dB hybrid couplers.

Advies Octrooi d.d. 1980-10-07	X AV	GV		B		BL
Opgave Mamo d.d. 1980-10-01	X AV	X GV	X SP	B		BL
Datum: - 1 OKT. 1980	Mamo					

DESIGN OF LINEAR POWER AMPLIFIERS FOR TV BANDS III-IV-V WITH THE BFQ68.

1. INTRODUCTION

The BFQ68 is a transistor intended for application in MATV amplifiers. It is however also very suitable for driver and output stages of TV transposers. The recommended operating point in class A is:

$V_{CE} = 15V, I_C = 240mA$

At a frequency of 860 MHz it will produce a peak sync output power of 780mW typ. at a 3-tone I.M. distortion of -60dB and a heatsink temperature of 25°C. The power gain is then 10,5 dB typ.

The high efficiency combined with the low supply voltage make the BFQ68 very suitable for application in transposers fed from solar cells.

The encapsulation is the ¼ inch stud type with ceramic cap (SOT-122).

In this report the theoretical designs of 3 wideband amplifiers with the BFQ68 will be given:

a. one for TV band IV-V (470-860 MHz)
b. one for TV band III (174-230 MHz) and
c. one for the combination of TV bands III, IV and V (174-860MHz)

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2. TRANSISTOR DATA

In the table below typical power gain, input impedance and optimum load impedance are given at 6 frequencies.

f(MHz)	G(dB)	$Z_i(\Omega)$	$Z_L(\Omega)$
174	23,9	6,36 - j2,52	45,9 + j6,9
202	22,7	6,30 - j1,59	45,5 + j7,9
230	21,6	6,26 - j0,81	45,0 + j8,9
470	15,5	6,13 + j3,58	39,2 + j15,6
665	12,6	6,07 + j6,24	33,5 + j18,4
860	10,5	6,00 + j8,67	28,1 + j19,3

Table I

A simplified equivalent circuit of the BFQ68 is shown in fig. 1.

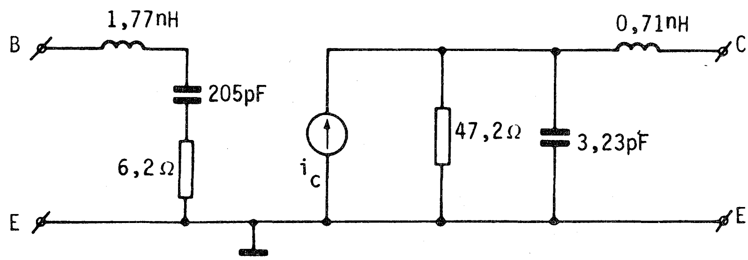

Fig. 1.

3. TV BAND IV-V AMPLIFIER

The circuit diagram is depicted in fig. 2.
The input network is the same as in Ref. 1.
The output network is a stripline version of the one described in Ref. 2.
The bias decoupling is only given for high frequencies.
The usual components for lower frequencies must of course be added (see e.g. Ref. 1.).

The following results have been obtained from calculations:

- the maximum V.S.W.R. of the output network is 1,08.
- the input V.S.W.R. varies from 1,43 (at 860 MHz) to 11,1 (at 470 MHz) to correct the variation of the power gain of the transistor.
- After combination of 2 amplifier branches by means of 3dB - 90° hybrid couplers *),the resulting power gain will be 10,1 ± 0,1 dB and the minimum peak sync output power for a 3-tone I.M. distortion of -60dB will be 1,38W.

  These data hold for typical transistors at a heatsink temperature of $25^{\circ}C$.

  *) See e.g. Ref. 4.

4. TV BAND III AMPLIFIER

A simular amplifier can be designed for use in TV band III (174-230MHz). The power gain is then of course much higher. Fig. 3 shows the schematic diagram. In the input network a damping resistor has been applied to ensure sufficient stability (R1).

The calculated performance is as follows:

- maximum V.S.W.R. of output network 1,01
- input V.S.W.R. varies from 1,39 (at 230 MHz) to 6,15 (at 174 MHz),
- After combining 2 amplifier branches with 3dB - 90° hybrid couplers, tne resulting power gain becomes 18,4 ± 0,1dB and the minimum peak sync output power for a 3-tone I.M. distortion of -60dB will be 1,48W.

5. TV BAND III-IV-V AMPLIFIER

Due to the favourable product of optimum load resistance and output capacitance of the BFQ68 amplifiers can be made with a very large bandwidth (up to 1GHz). Fig. 4 shows the schematic diagram of an amplifier that covers the TV bands III, IV and V (174 - 860MHz).

In the input network a resistor (R1) has been added across the 50 ohms input terminals to limit the input V.S.W.R. in band III.

The output network is a 6 elements Chebyshev bandpass filter as described in Ref. 1.

Calculation of the performance shows the following results:

- maximum V.S.W.R. of output network is 1,16
- input V.S.W.R. changes from 1,72 at 860 MHz to 10,6 at 174 MHz
- combination of 2 amplifier branches with 3dB -90^{o} hybrid couplers yields a peak sync output power of 1,29 W min. at a 3-tone I.M. distortion of -60dB.
  The power gain is then 9,7 $\pm$ 0,2dB.

A disadvantage of the system described here is that hybrid couplers covering the whole frequency range from 174-860MHz are not available. Therefore these couplers have to be changed when going from band III to IV/V or vice versa.

6. DRIVER AMPLIFIER

The amplifier described in Ref. 3 is very suitable for driving the amplifier stages described in this report. It is a cascade connection of BFQ34 and BFQ68 amplifiers with a power gain of 15,5 $\pm$ 0,3dB in the frequency range of 40 - 860 MHz.
The peak sync output power for a 3-tone I.M. distortion of -60dB is 150mW in bands IV/V and 285mW min. in band III.
Table II gives a survey of the calculated overall performance when the driver of Ref. 3. is combined with one of the 3 amplifiers described in this report, taking into account that 2 branches are coupled by means of hybrids.

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Driver	BFQ34 + BFQ68 (Ref. 3)		
Output stage	2 x BFQ68 (Fig.2)	2 x BFQ68 (Fig.3)	2 x BFQ68(Fig. 4)
Freq. range (MHz)	470 - 860	174 - 230	174 - 860
Power gain (dB)	25,6 $\pm$ 0,4	33,9 $\pm$ 0,4	25,2 $\pm$ 0,5
I.M. distortion (dB) at $P_o = 1W_{ps}$	-57	-63	-56
X-mod. (%) at $P_o = 1W_{ps}$	5	2,5	5,6

Table II

REFERENCES

1. A.H. Hilbers and M.J. Köppen, "Wide-band linear power amplifiers (470-860MHz) with the transistors BLW32 and BLW33", C.A.B. report nr. ECO 7806.
2. A.H. Hilbers, "Design of a linear power amplifier for TV band III with the BLV30", C.A.B. report nr. ECO 8004.
3. G. Lukkassen, "A wide-band amplifier with the BFQ34 and BFQ68", Appl. report nr. NCO 8003.
4. M.J. Köppen, "Wideband linear power amplifier (470 - 860 MHz) with two transistors BLW98", C.A.B. report nr. ECO 7905.

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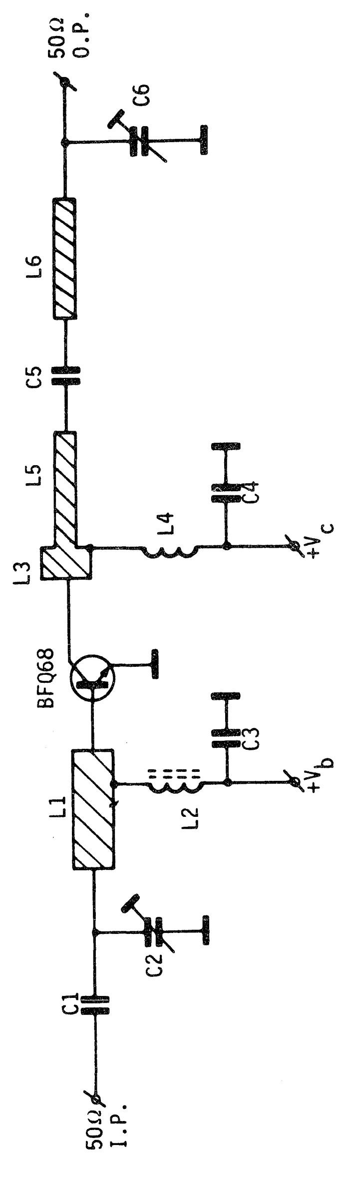

FIG. 2
Band IV - V amplifier

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Parts list Fig. 2

C1 = 5,6pF, ceramic chip

C2 = parallel connection of 2,2pF, ceramic chip and 1,2 - 3,5 pF, film dielectric trimmer.

C3 = 68pF, ceramic chip

C4 = 68pF, ceramic chip

C5 = 13pF, ceramic chip

C6 = 1,2 - 3,5pF, film dielectric trimmer

L1 = 38Ω stripline; W = 6mm, l = 10,6 mm.

L2 = 1 μH, microchoke

L3 = 38Ω stripline; W = 6mm, l = 3,0mm.

L4 = 18,9nH; 2 turns of 0,7mm copperwire, int. diam. = 3mm, length = 2,1mm, leads: 2 x 5 mm

L5 = 86Ω stripline; W = 1,5mm, l = 11,2mm

L6 = 86Ω stripline; W = 1,5mm, l = 11,2mm

P.C. board material: 1/16" teflon fibre-glass (ε_r = 2,74)

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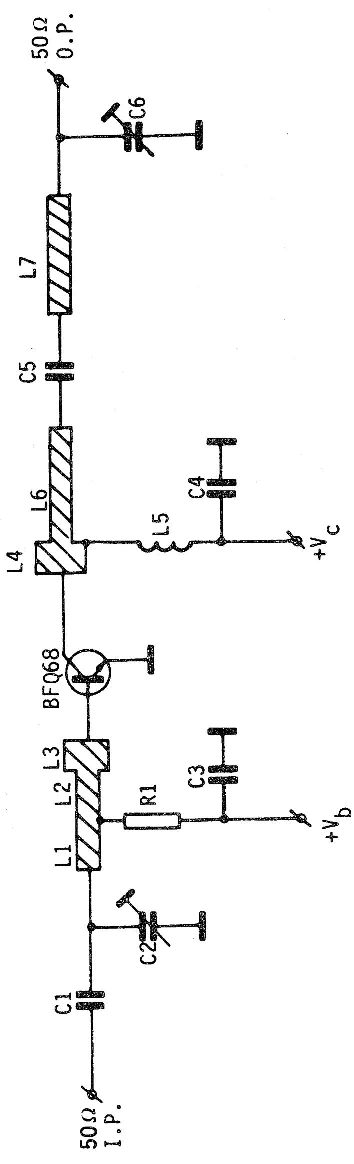

FIG.3

Band III amplifier

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Parts list Fig. 3

R1 = 10Ω, carbon

C1 = 24pF, ceramic chip
C2 = parallel connection of 18pF, ceramic chip and
4 - 40pF, film dielectric trimmer
C3 = 680pF, ceramic chip
C4 = 680pF, ceramic chip
C5 = 91pF, ceramic chip
C6 = 1,2 - 3,5pF, film dielectric trimmer

L1 = 70Ω stripline; w = 1,5mm, l = 15,5mm
L2 = 70Ω stripline; w = 1,5mm, l = 16,5mm
L3 = 30Ω stripline; w = 6mm , l = 3,0mm
L4 = 30Ω stripline; W = 6mm, l = 3,0 mm
L5 = 159nH; 8 turns of 0,5 mm copper wire,
int. diam. = 4,5mm, length = 8,0mm, leads: 2 x 5 mm.
L6 = 70Ω stripline; w = 1,5mm, l = 24,1 mm
L7 = 70Ω stripline; w = 1,5mm, l = 24,1mm

P.C. board material: 1/16" epoxy fibre-glass ($\varepsilon_r \simeq 4{,}5$)

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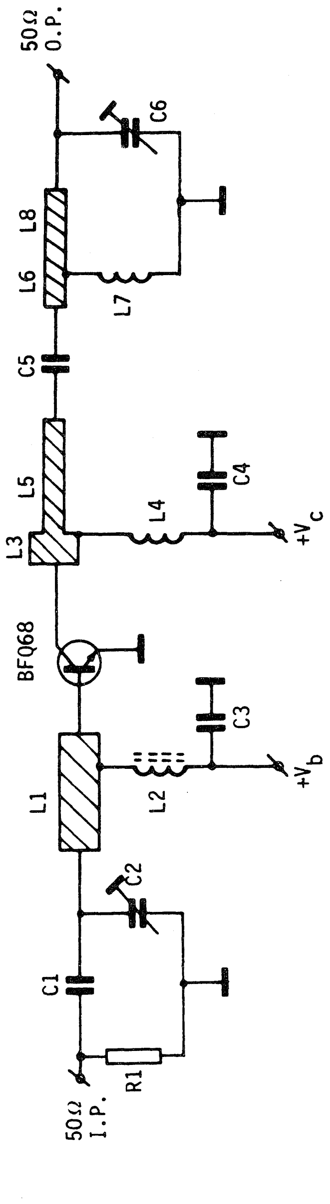

FIG. 4
Band III-IV-V amplifier

Electronic components and materials

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Parts list Fig. 4

R1 = 560Ω, carbon

C1 = 6,2pF, ceramic chip

C2 = parallel connection of 2,4pF, ceramic chip and 1,2 - 3,5pF, film dielectric trimmer

C3 = parallel connection of 270pF and 100pF, ceramic chips.

C4 = parallel connection of 270pF and 100pF, ceramic chips.

C5 = 16pF, ceramic chip

C6 = parallel connection of 1,5pF, ceramic chip and 0,5 - 2,0pF, Micro thin trim capacitor (Tekelec-Airtronic)

L1 = 38Ω stripline; w = 6mm, l=9,5mm

L2 = 1μH, microchoke

L3 = 38Ω stripline; w = 6mm, l = 3,0mm

L4 = 43,1nH; 5 turns of 0,5mm copper wire, int. diam. = 2,5mm, length = 4,5mm, leads: 2 x 5mm

L5 = 86Ω stripline; w = 1,5mm, l = 6,3mm

L6 = 86Ω stripline; w = 1,5mm, l = 6,3mm

L7 = 48,2nH; 5 turns of 0,5mm copper wire, int. diam. = 3,0mm, length = 5,5mm, leads: 2 x 5mm

L8 = 86Ω stripline; w=1,5mm, l = 10,2mm

P.C. board material: 1/16" teflon fibre-glass (ϵ_r = 2,74)

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NIJMEGEN – THE NETHERLANDS

REPORT No: NCO 8004
PROJECT No:
AUTHOR: H. van Hees
DATE: 26.09.1980

TITLE

A wide-band linear power amplifier (174-230MHz) with two transistors BLV 32 F

ABSTRACT

For application in driver or final stages of TV transposers in band III (174-230MHz) a linear wide-band power amplifier has been designed with two transistors BLV 32 F, coupled by means of 3dB-90° hybrids. The class-A D.C. setting of the transistors is $I_E = 1.5A$ and $V_{CE} = 28V$.

Two prototypes have been built, the main properties of which are:

174-230MHz	amplifier I	amplifier II	
P_{osync} at -55dB intermodulation (three tone)	23.8-26.0	22.5-27	W
cross modulation at P_o = 20 Watt	7.5-9.5	6.5-10	%
power gain at P_o = 20 Watt	15.9 ± 0.3	16.1 ± 0.3	dB
return losses at input and output	≥ 23	≥ 21.5	dB

appr. J. Tuil

Advice Patents Dept. d.d: 30-10-80	AV X	GV		B		BL
Decision MAMO d.d: 13-10-80	AV X	GV X	EI X	B		BL
DATE: 1980-10-13	MAMO:					

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NIJMEGEN – THE NETHERLANDS

REPORT No: NCO 8004	AUTHOR: H. van Hees
PROJECT No:	DATE: 26.09.1980

TITLE

A wide-band linear power amplifier (174-230MHz) with two transistors BLV 32 F

SUMMARY

For application in TV transposers a wide-band linear power amplifier has been designed with two transistors BLV 32 F. It is a hybrid-coupled amplifier for the frequency range of 174-230MHz.
For this linear application the transistors are operated in class A. The D.C. setting is I_E = 1.5A and V_{CE} = 28V.
The circuit has been designed according to ref. 1 and ref. 2 and after that submitted to a computer optimization program.
Fig. A shows the circuit of the amplifier of which two prototypes have been built. The applied circuit board is a double copper clad print with 1/16" fibre glass as a dielectric ($\varepsilon_r \approx 4.5$).
Practical optimization and tuning has been done on a dynamic gain compression set-up.
Fig. B gives the small signal gain and return losses of both amplifiers.
The output power P_{osync} at three intermodulation levels is given in fig. C.
At P_o = 20 Watt the cross modulation of amplifier I varies between 7.5% and 9.5% and of amplifier II between 6.5% and 10% (see fig. D).

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	DATE: 1980-10-13	MAMO:					

Fig. E and F show the gain at a constant output power P_o = 20 Watt and indicate that the 1dB gain compression occurs at $P_o \simeq$ 46 Watt for amplifier I and at $P_o \simeq$ 43 Watt for amplifier II.
The heatsink has forced air cooling and the temperature remains below 40°C at an ambient temperature of 25°C.
The supply voltage V_S = 30V.

Ref. 1: G. Matthaei
Tables of Chebychev impedance transforming networks of low-pass filter form.
Proc. of the IEEE, August 1964.

Ref. 2: O. Pitzalis Jr. and R.A. Gibson
Tables of impedance matching networks which approximate prescribed attenuation versus frequency slopes.
IEEE transactions on microwave theory and techniques, vol. MTT 119, no. 4, April 1971.

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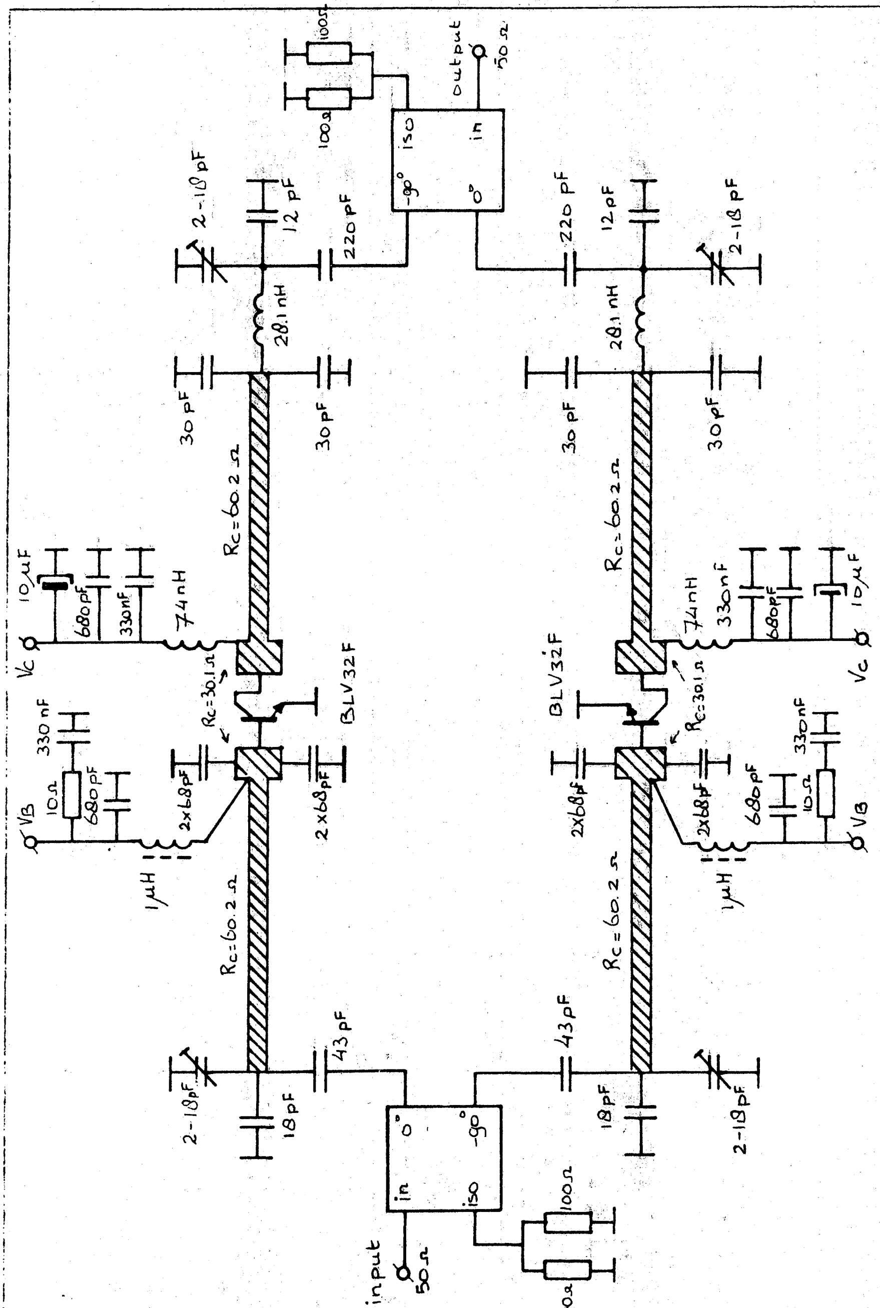

Fig.A Circuit diagram

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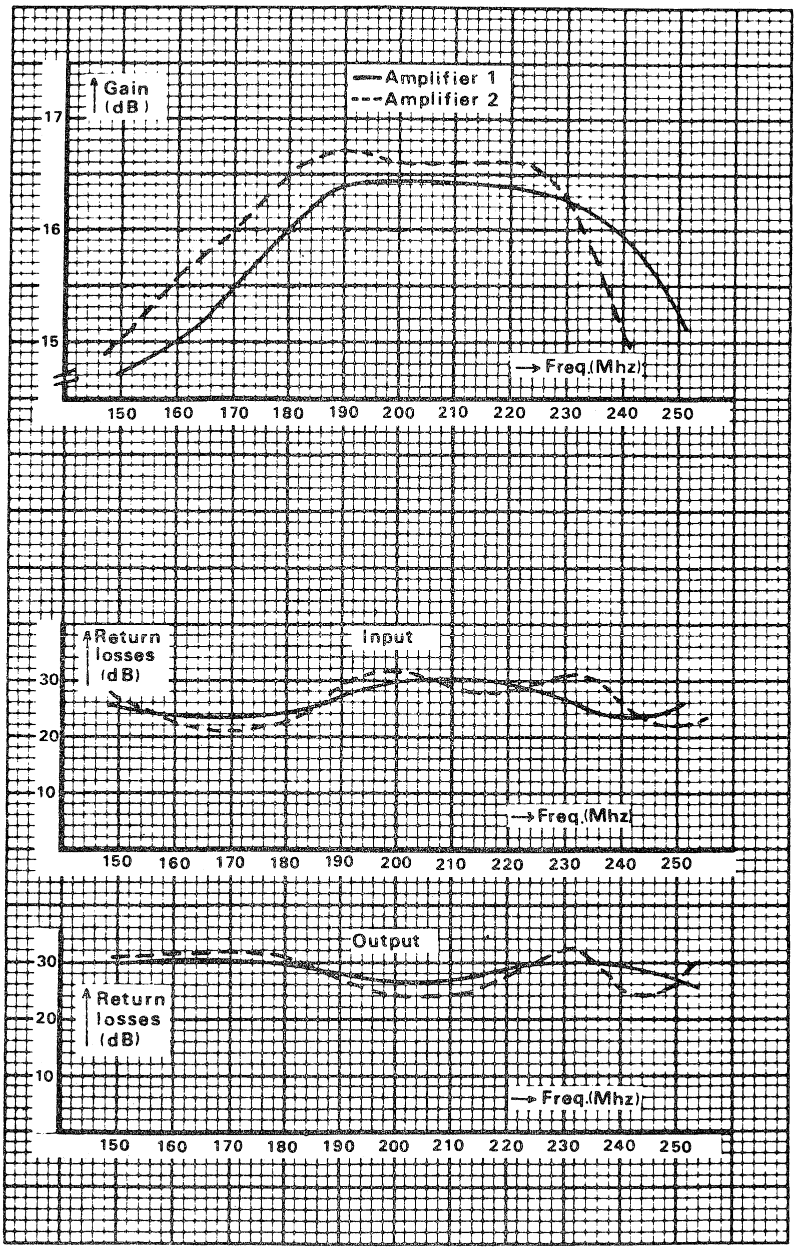

Fig. B Small signal gain and return losses

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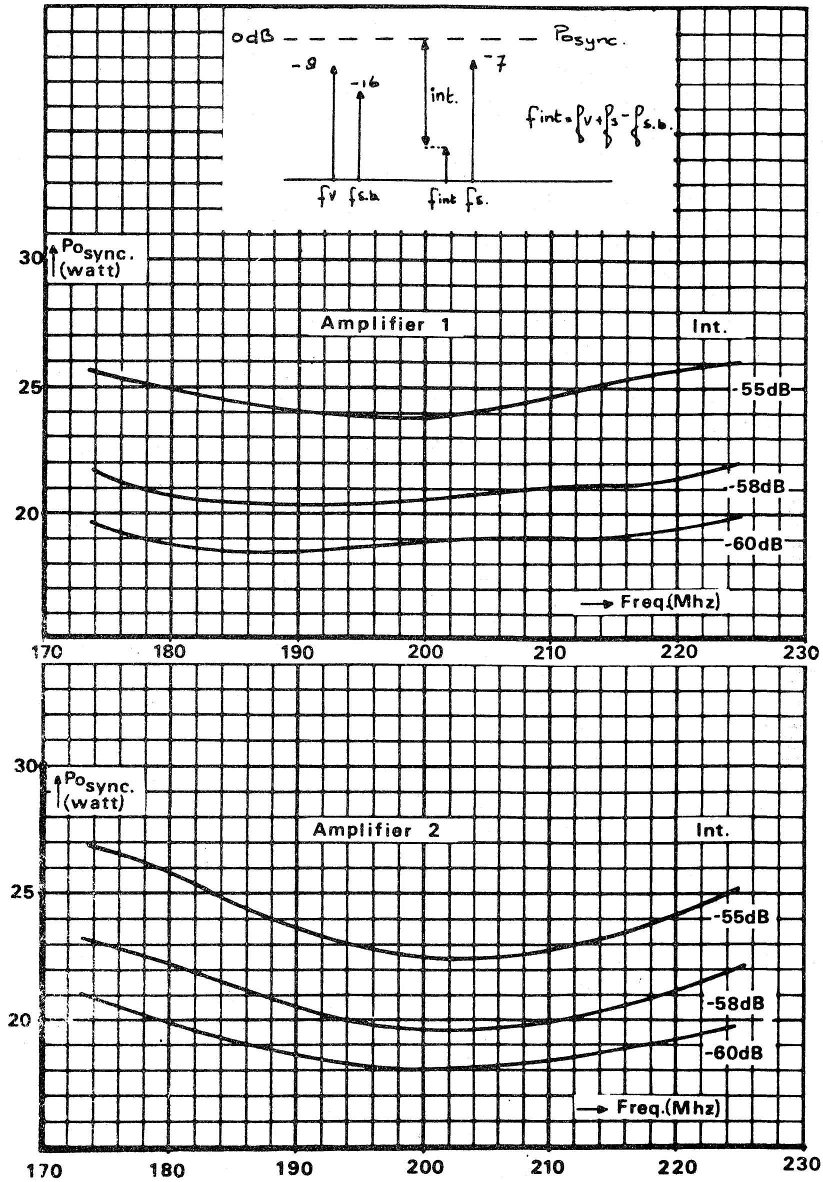

Fig. C Output power $P_{o_{sync}}$

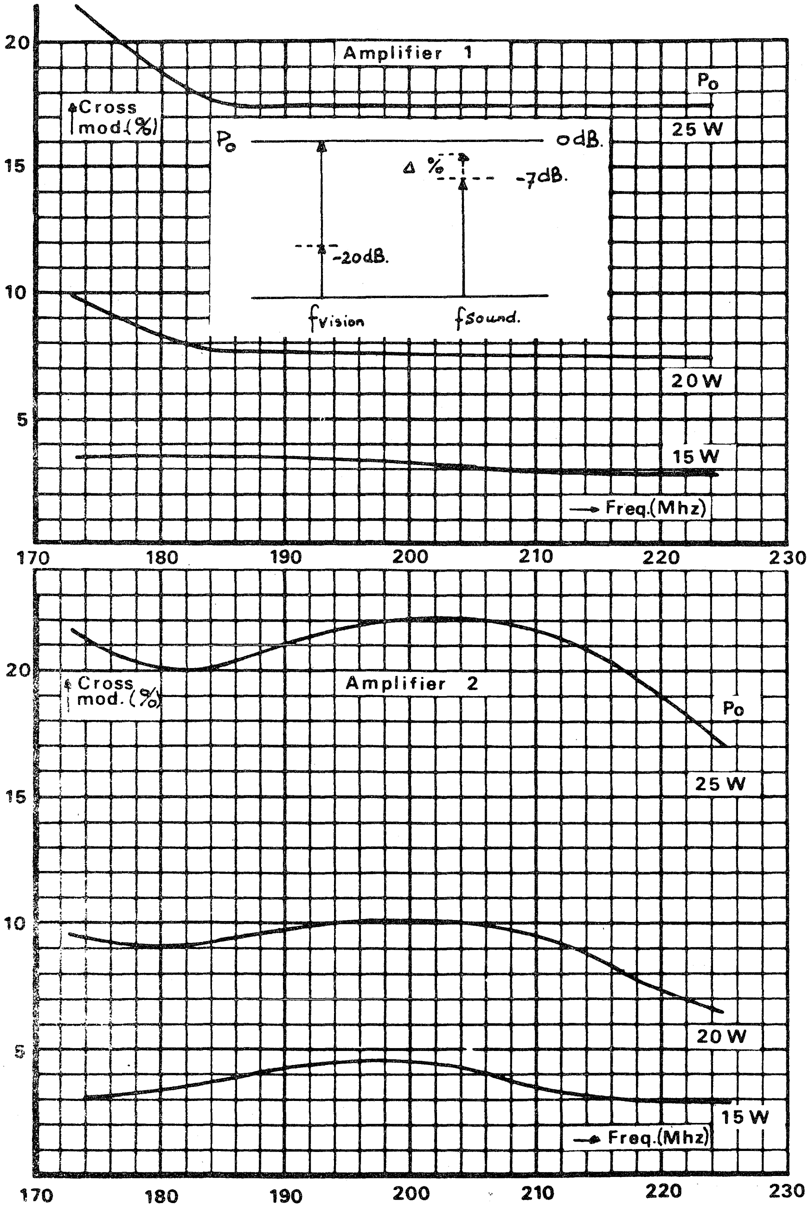

Fig.D Cross modulation

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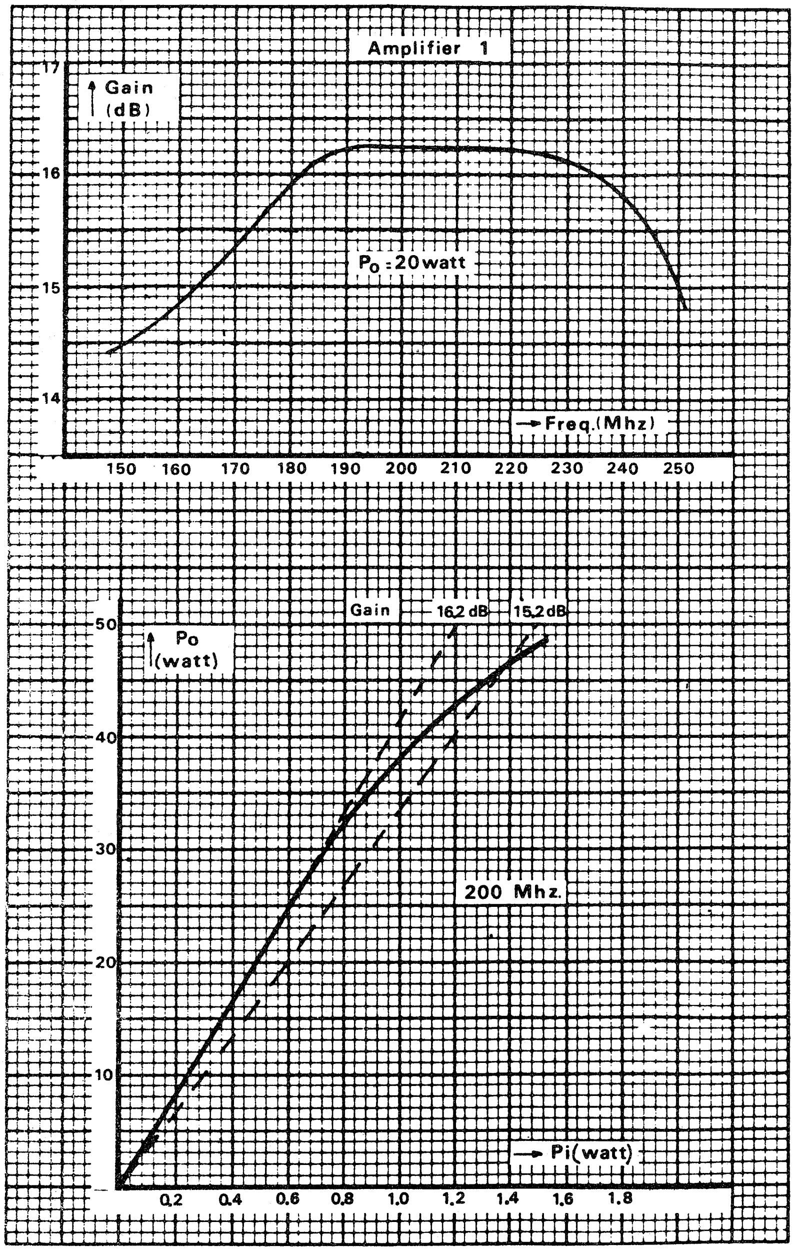

Fig.E Gain and gain compression

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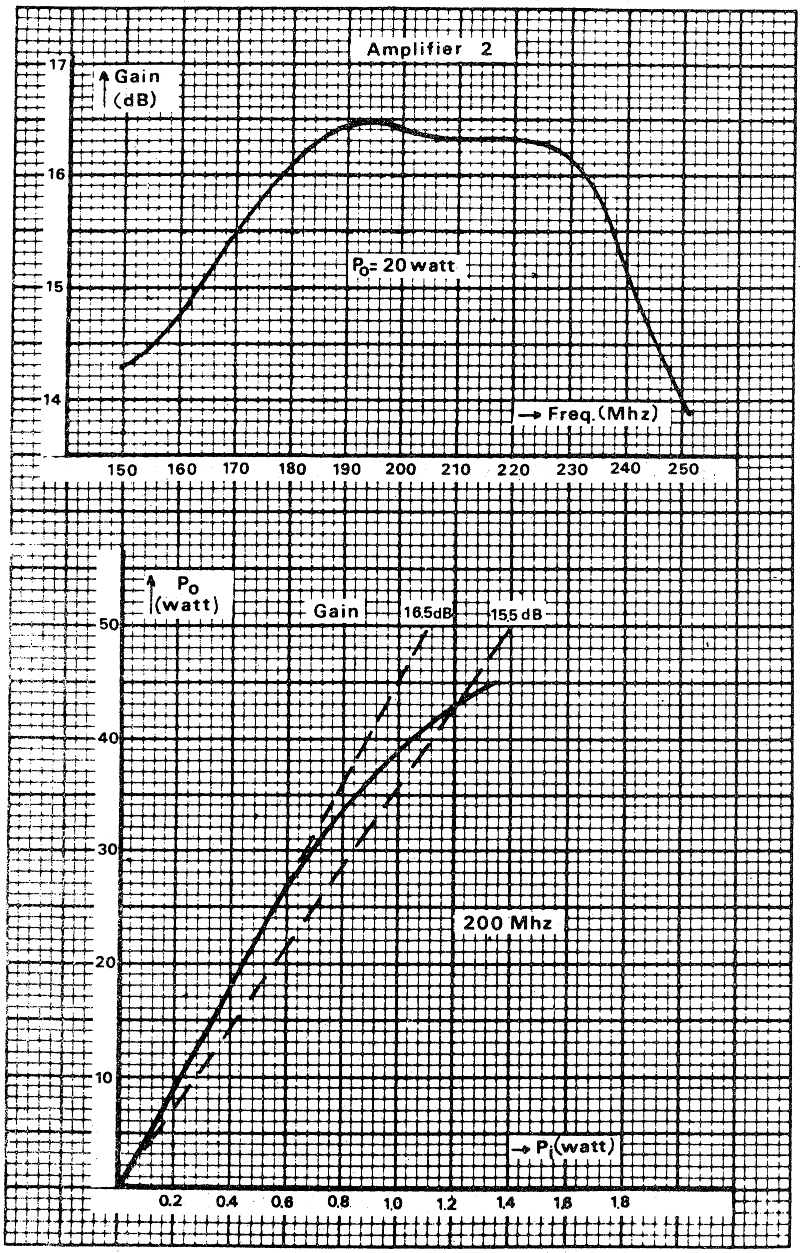

Fig. F Gain and gain compression

SEMICONDUCTOR APPLICATION LABORATORY S.A.L.NIJMEGEN-THE NETHERLANDS

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1. Introduction

For application in TV transposers for band III (174-230MHz) a wide-band linear power amplifier has been designed with two transistors BLV 32 F.

The D.C. adjustment of these transistors is I_C = 1.5A and V_{CE} = 28V.

The encapsulation is a 3/8" six leads flange envelope with a ceramic cap (NO 220).

2. Design of the amplifier

2.1 Some properties of the BLV 32 F

For class A operation the BLV 32 F is specified at V_{CE} = 28V and I_C = 1.5A.

The typical gain, input- and load impedance are given below in table 1.

freq. (MHz)	gain (dB)	input impedance (Ω)	load impedance (Ω)
174	18.73	0.65 + j 1.18	8.16 + j 7.47
202	17.42	0.65 + j 1.46	6.89 + j 7.15
230	16.26	0.65 + j 1.73	5.82 + j 6.70

table 1

2.2 Output network

The 50Ω system impedance has to be transformed into the optimum load impedance for the transistor, which is given in table 1. This is done by means of an L-C output network.

The circuit has been calculated according to ref. 1 and submitted to a computer optimization program.

Fig. 1 shows the result.

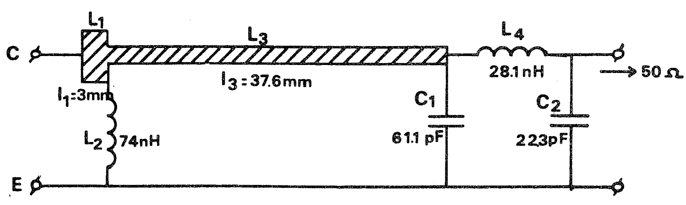

Fig. 1 Output circuit.

It starts with a stripline, width 6mm, which is the soldering place for the collector lead of the transistor. The collector choke L_2, for the D.C. biasing, is connected to the stripline at a distance of 3mm from the transistor.
The applied p.c. board is 1/16" glass-fibre with an $\varepsilon_r \simeq 4.5$.
L_3 is a stripline, width 2mm and length 37.6mm.
C_1 consists of 2 chip capacitors in parallel and C_2 one chip capacitor in parallel with a film dielectric trimmer.

2.3 Input network

In the frequency range of 174-230MHz the power gain of the BLV 32 F varies about 2.5dB (see table 1). The gain slope can be decreased by applying appropriate mismatch at the lower frequencies. The increasing insertion loss of the network approximately compensates the increasing power gain. This method is described in ref. 2.
Fig. 2 shows the calculated input circuit after computer optimization.

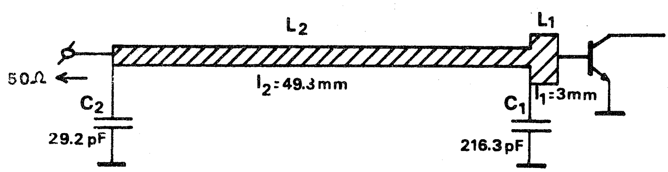

Fig.2 Input circuit.

The inductances L_1 and L_2 are executed as striplines. L_1, width 6mm to be able to solder the base lead of the transistor; L_2 has a width of 2mm. C_1 consists of two chip capacitors in parallel and C_2 is one chip capacitor in parallel with a film dielectric trimmer.

3. The hybrid coupled amplifier

3.1 Practical considerations

In the previous section the theoretical design of a single stage wide-band BLV 32 F amplifier has been discussed.

The gain is made sufficiently flat at the cost of an impermissably high input VSWR. Therefore, in practice, two wide-band stages are coupled in parallel by means of two wide-band 3dB-90^{o} coaxial hybrids. The properties of these hybrids reduce the input VSWR to about 1.2. The reflected power of the single stages is absorbed by the 50Ω resistance at the isolated port. Fig. 3 shows the circuit diagram of the complete 2x BLV 32 F amplifier. The amplifier has been designed on a printed circuit board with epoxy fibre glass as dielectric ($\varepsilon_r \simeq 4.5$). The thickness of the board is 1/16" and it is copper clad on both sides. Fig. 4 shows a positive copy of this p.c. board. To get a good contact between upper and lower side, rivets have been used at several places. For the same reason, at the edges of the board, copper straps have been soldered.

Fig. 5 is the amplifier lay-out.

Each transistor has his own bias unit to obtain a stable D.C. setting. Fig. 6 gives the circuit of such a bias unit. The supply voltage for the amplifier V_S = 30V.

3.2 Power sweep

We have built two prototypes of this design. The amplifiers have been optimized by means of a power sweep set-up. With this set-up it is possible to measure gain compression and output power under swept conditions. Gain compression of an amplifier correlates with the distortion. So with this power sweep measurement it was possible to find the best compromise between a sufficiently flat gain and a high output power at a low distortion level. This optimum has been established by trimming the capacitors C_5, C_6, C_{25} and C_{26}. Moreover, the base-emitter capacitances C_7 till C_{10} had to be increased and shifted to the midst of L_3 and L_4.

4. Measured performance

4.1 Small signal gain and return losses

Fig. 7 shows the gain and return losses as a function of frequency, measured under small signal conditions.

Results:

174-230MHz	amplifier I		amplifier II		
	min.	max.	min.	max.	
gain	15.7	16.4	16.4	16.7	dB
return losses input	24	30	21	31	dB
return losses output	26	30	24	32	dB

4.2 Intermodulation

In fig. 8 is the output power at three intermodulation levels. It has been measured according to the three-tone test method (vision carrier -8dB, sound carrier -7dB, and side-band carrier -16dB). Zero dB corresponds to the peak sync. level.

The minimum output power at -55dB intermodulation amounts to 23.8 Watt for amplifier I and 22.5 Watt for no. II.

4.3 Cross modulation

Fig. 9 shows the cross modulation through the band at three P_o levels. It is a two tone measurement (vision carrier 0dB, sound carrier -7dB). The amplitude of the vision carrier is changed from white level (-20dB) to peak sync. level (0dB). The observed change of the voltage amplitude of the sound carrier is called cross modulation. It is expressed as a percentage of the amplitude of the sound carrier (vision carrier at white level → -20dB).

It is measured with the spectrum analyzer; which is operated in the linear mode.

At a P_o level of 20 Watt the cross modulation of amplifier I varies from 7.5% to 9.5% and of amplifier II from 6.5% to 10%.

4.4 Gain and gain compression at high output power

The gain versus frequency at a constant output power of 20 Watt is shown in fig. 10 and fig. 11.

From the measured P_o versus P_i curves one can see that at 200MHz the 1dB gain compression occurs at an output power of about 46 Watt for amplifier I and 43 Watt for amplifier II.

5. Conclusion

It is possible to build a linear amplifier with excellent performance with two transistors BLV 32 F.

The main properties of two prototypes are:

band III		amplifier I	amplifier II
gain (P_o = 20 Watt)	dB	15.9 $\pm$0.3	16.1 $\pm$0.3
return losses at input and output	dB	$\geqslant$ 23	$\geqslant$ 21
P_o at 1dB gain compression (200MHz)	W	46	43
P_{osync} at -55dB inter-modulation	W	$\geqslant$ 23	$\geqslant$ 21.5
cross modulation at P_o = 20 Watt	%	$\leqslant$ 9.5	$\leqslant$ 10

6. References

Ref. 1: G.L. Matthaei
Tables of Chebychev impedance transforming networks of low-pass filter form.
Proc. of the IEEE, August 1964.

Ref. 2: O. Pitzalis Jr. and R.A. Gibson
Tables of impedance matching networks which approximate prescribed attenuation versus frequency slopes.
IEEE transactions on microwave theory and techniques, vol. MTT 119, no. 4, April 1971.

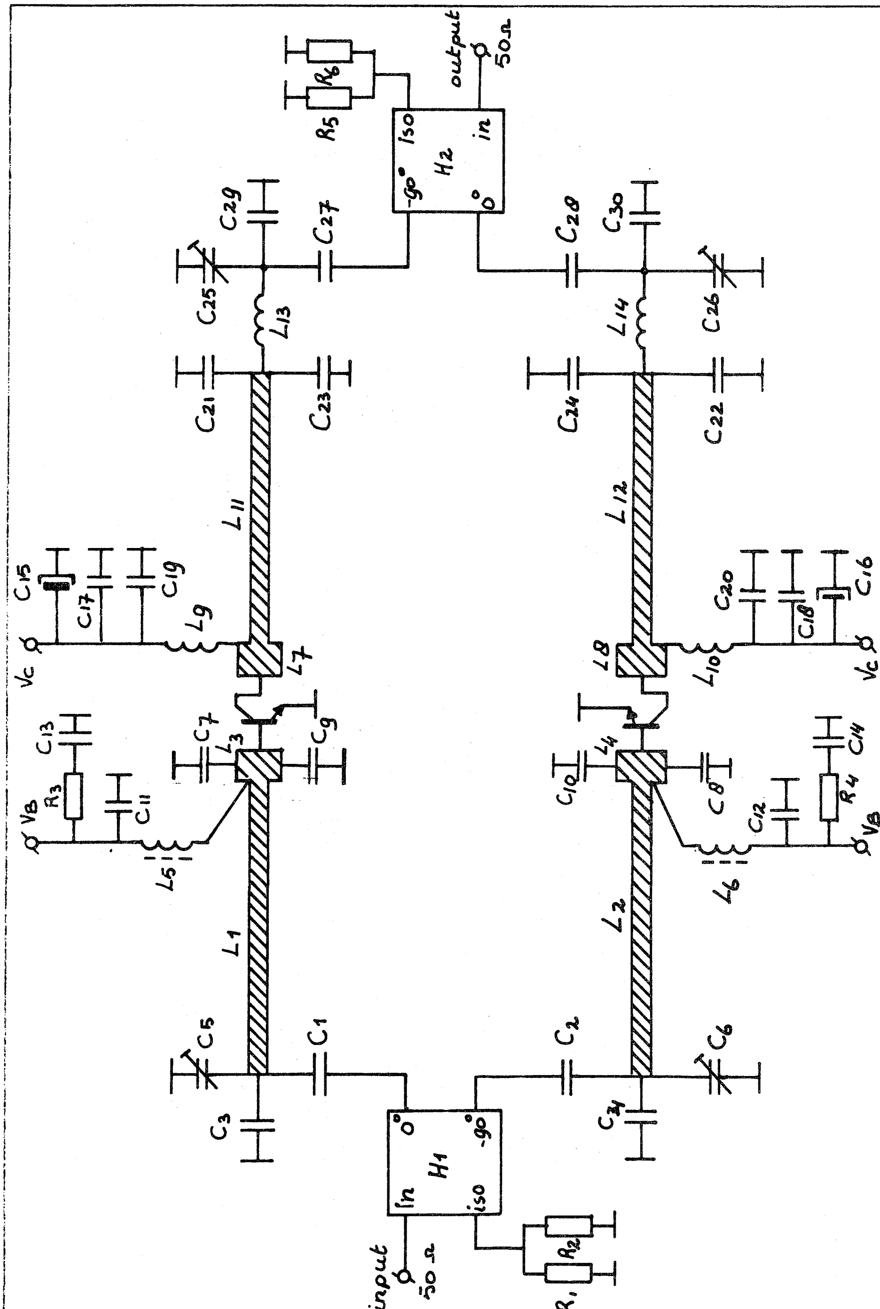

Fig.3 Circuit diagram.

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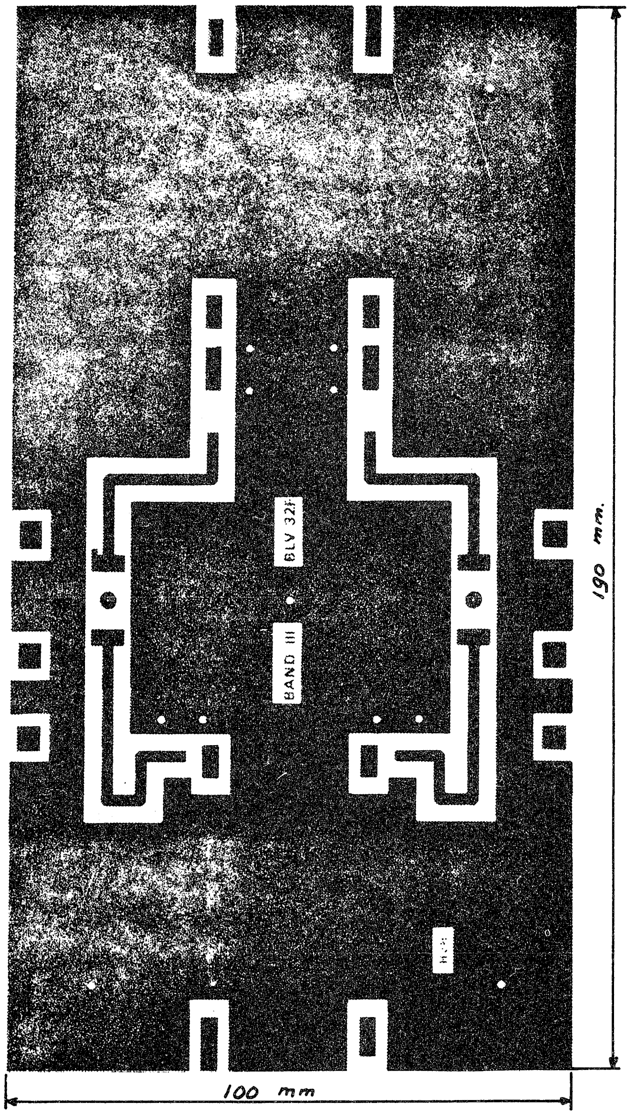

Fig. 4 P.C. board.

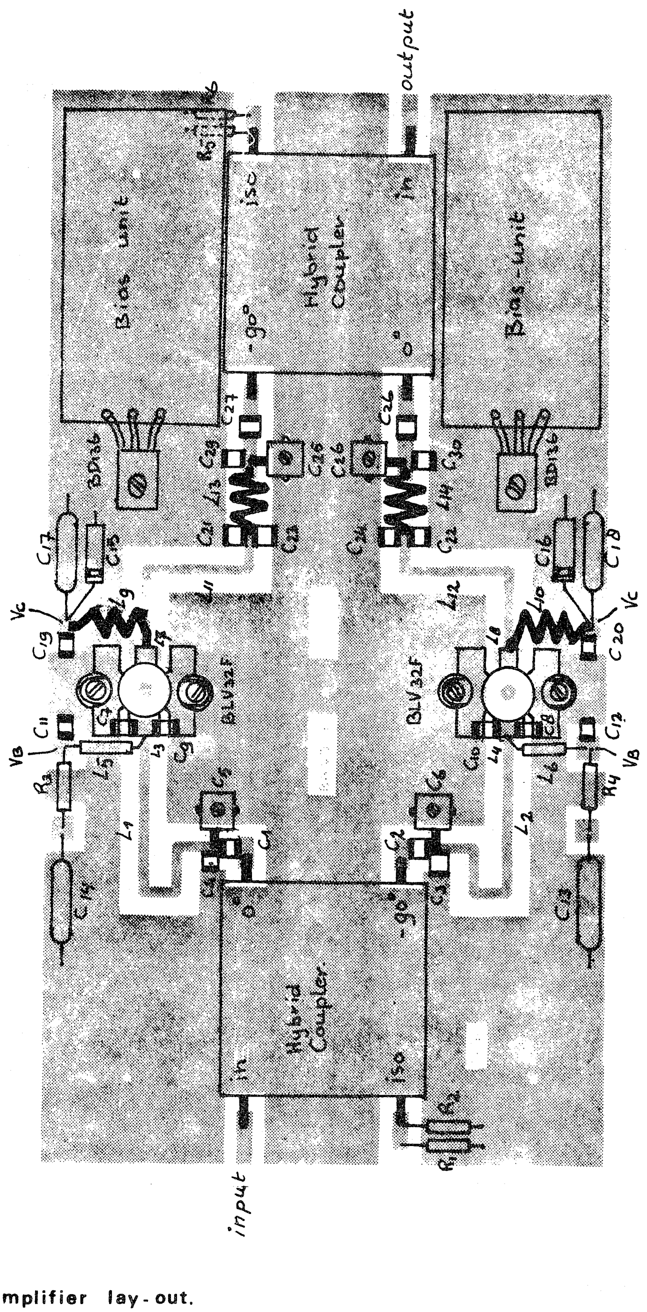

Fig.5 Amplifier lay-out.

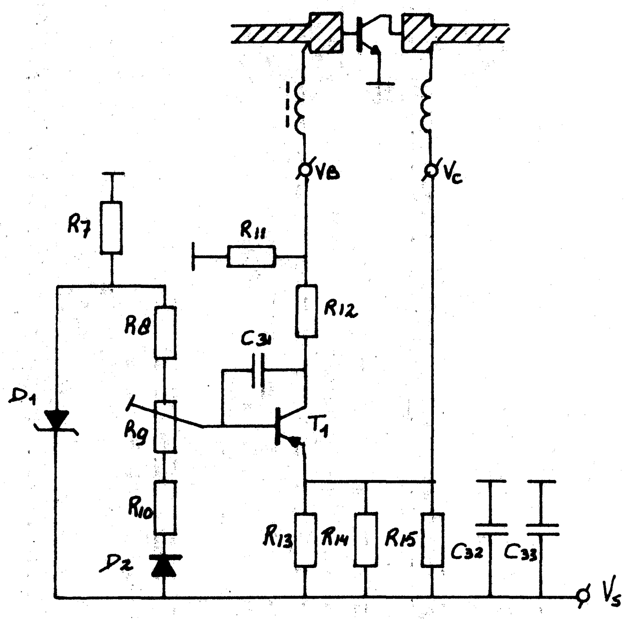

Fig.6 Bias circuit.

7. List of components

$C_1 = C_2$ = 43pF chip capacitor, ATC 100 B.

$C_3 = C_4$ = 18pF chip capacitor, ATC 100 B.

$C_5 = C_6 = C_{25} = C_{26}$ = 1.8-18pF, film dielectric trimmer. cat no. 2222 809 09003.

$C_7 = C_8 = C_9 = C_{10}$ = 2x 68pF chip capacitor, ATC 100 B.

$C_{11} = C_{12} = C_{19} = C_{20}$ = 680pF chip capacitor, Philips NPO cat no. 2222 852 13681.

$C_{13} = C_{14} = C_{17} = C_{18}$ = 330nf, metalized film capacitor.

$C_{21} = C_{22} = C_{23} = C_{24}$ = 30pF chip capacitor, ATC 100 B.

$C_{27} = C_{28}$ = 220pF chip capacitor, ATC 100 B.

$C_{29} = C_{30}$ = 12pF chip capacitor, ATC 100 B.

$C_{31} = C_{32}$ = 10kpF ceramic capacitor, 63V. cat no. 2222 629 01103.

$C_{15} = C_{16}$ = 10µF, 63V electrolytic capacitor.

C_{33} = 100nF, metalized film capacitor.

$R_1 = R_2 = R_5 = R_6$ = 100Ω metal film resistor PR 37.

$R_3 = R_4$ = 10Ω carbon film resistor CR 25.

R_7 = 1k6 carbon film resistor CR 37.

R_8 = 390Ω carbon film resistor CR 25.

R_9 = 220Ω cermet potentiometer.

$R_{10} = R_{11}$ = 150Ω carbon film resistor CR 25.

R_{12} = 110Ω carbon film resistor CR 37.

$R_{13} = R_{14} = R_{15}$ = 3.9Ω carbon film resistor CR 68.

$L_1 = L_2$ = 60.2Ω stripline w = 2mm l = 49.3mm.

$L_3 = L_4 = L_7 = L_8$ = 30.1Ω stripline w = 6mm l = 3mm.

$L_5 = L_6$ = 1µH choke

$L_9 = L_{10}$ = 74nH 6 turns of 1mm enamelled Cu wire. D_{int} = 4mm l = 10.4mm.

$L_{11} = L_{12}$ = 60.2Ω stripline w = 2mm l = 40.4mm.

$L_{13} = L_{14}$ = 28.1nH 3 turns of 1mm enamelled Cu wire. D_{int} = 4mm l = 7mm.

$H_1 = H_2$ = 3dB-90° hybrid coupler, model no. 10262-3 Anaren Microwave Inc.

D_1 = BZX 79-C6V8.

D_2 = BY 206.

T_1 = BD 136.

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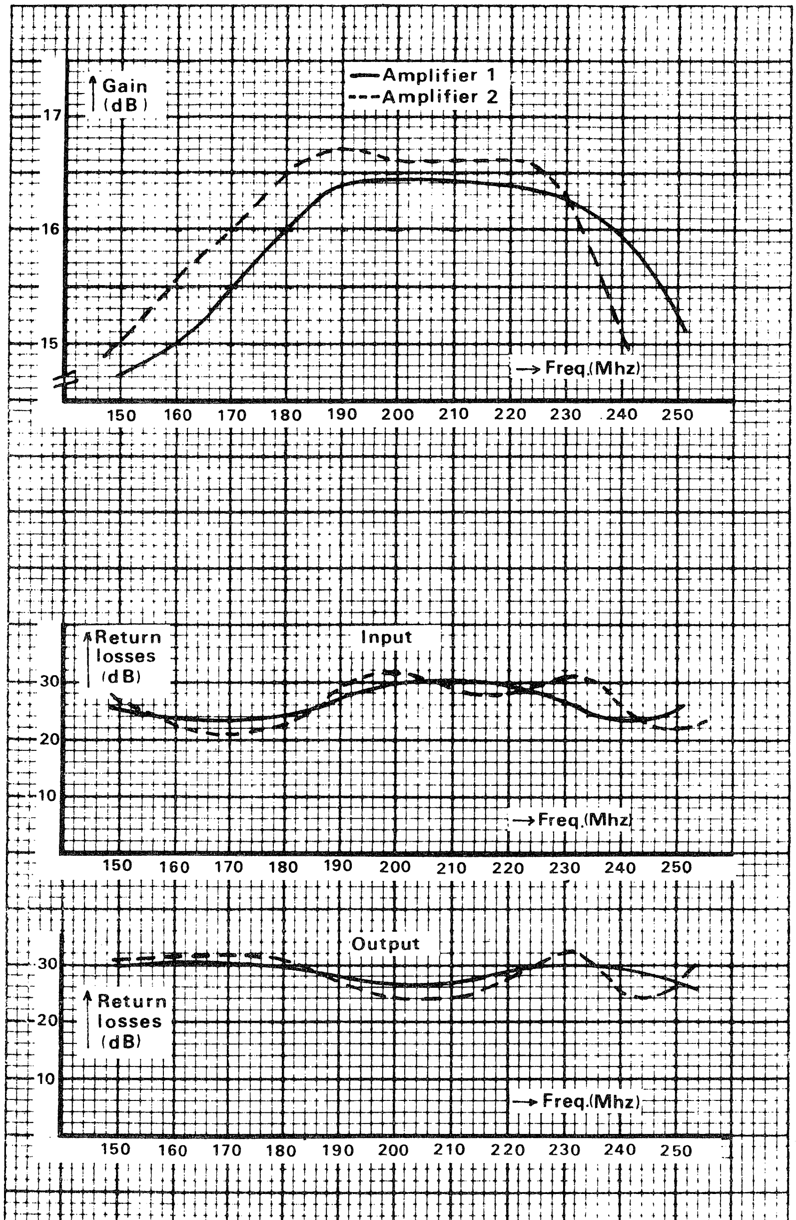

Fig.7 Small signal gain and return losses.

Electronic components and materials

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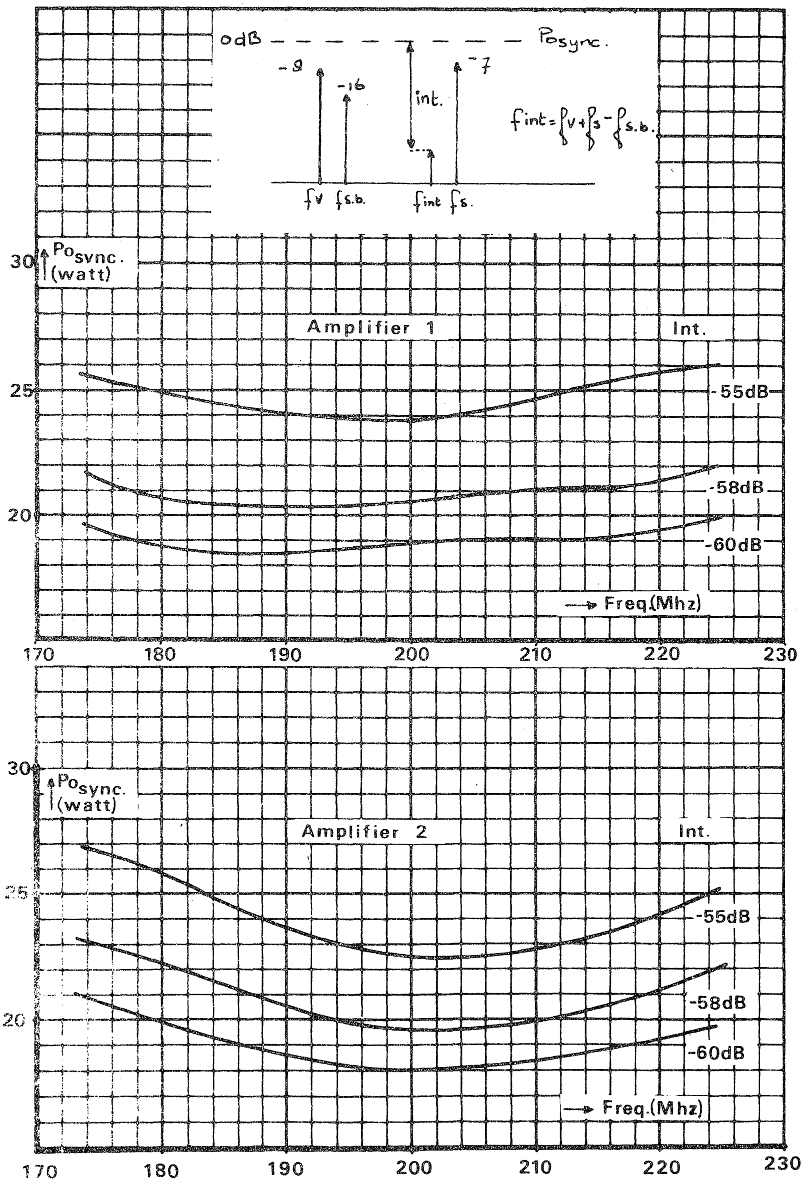

Fig.8 Output power $Po_{sync.}$

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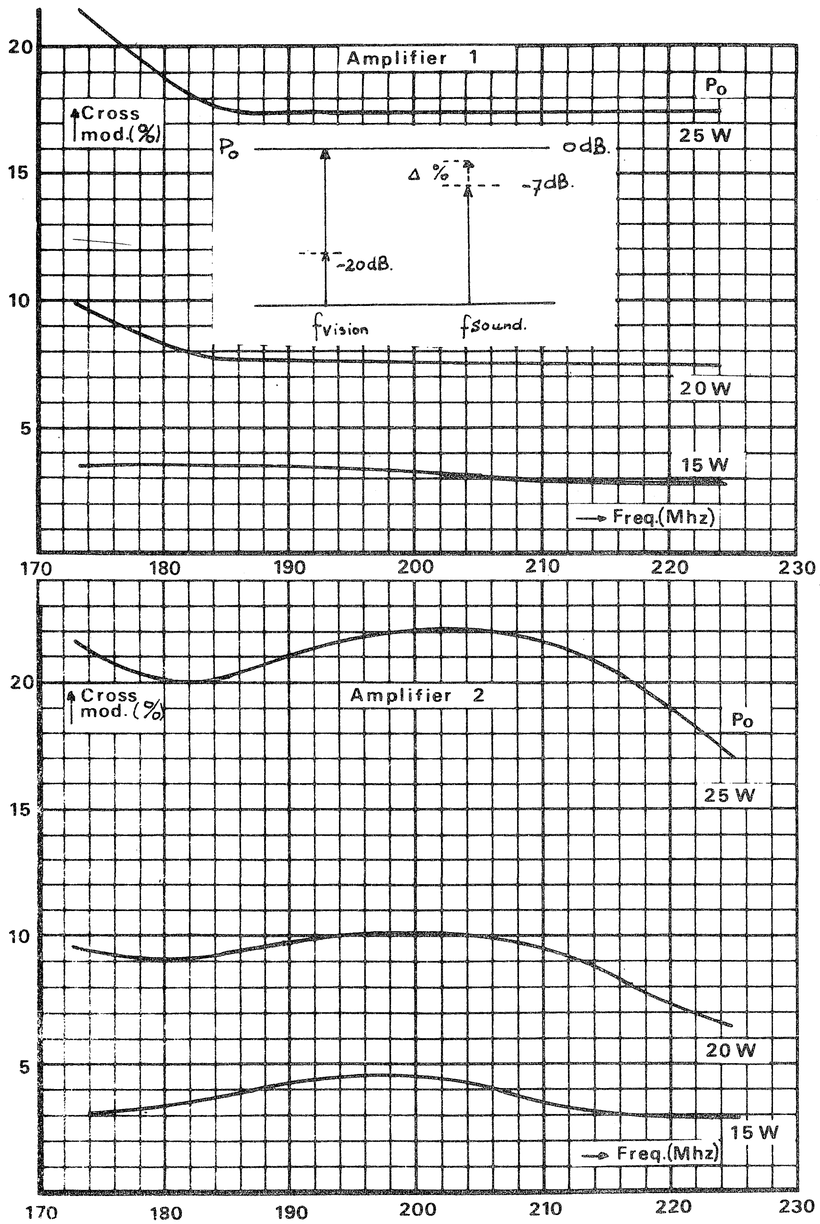

Fig.9 Cross modulation.

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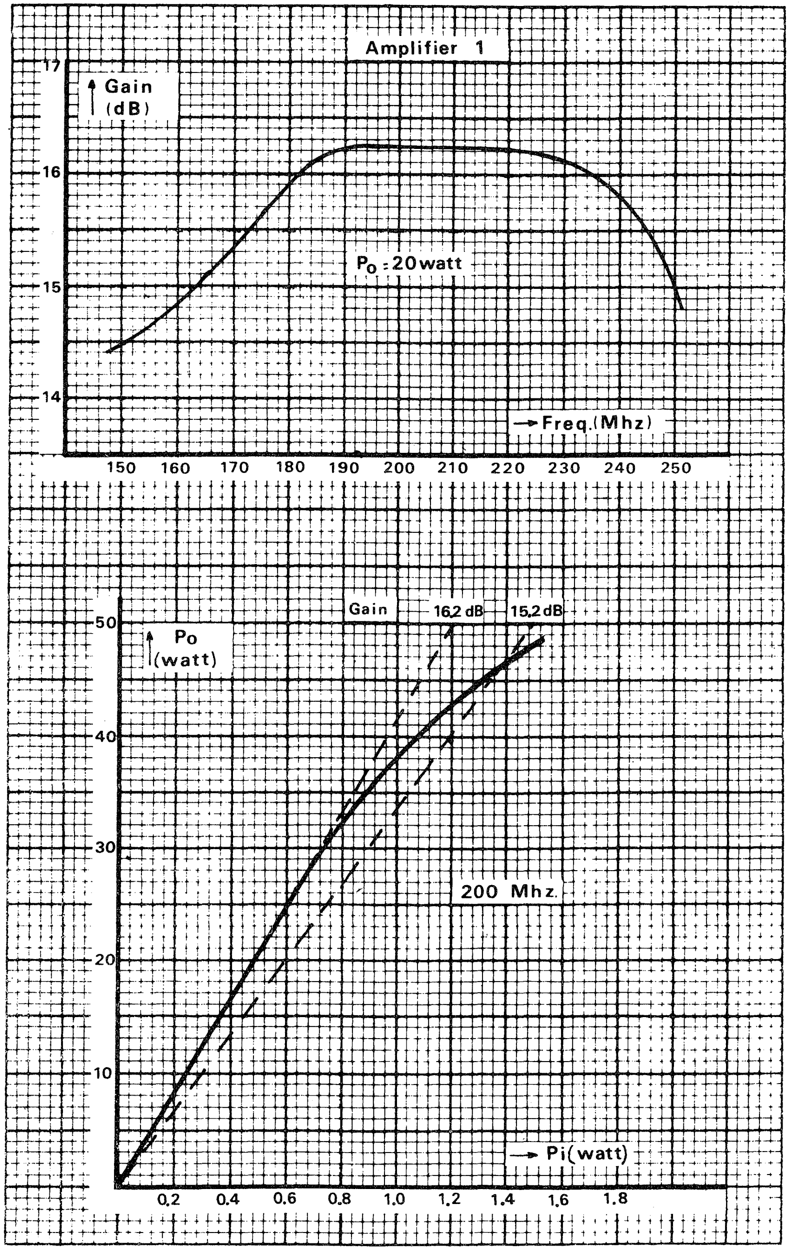

Fig.10 Gain and gain compression.

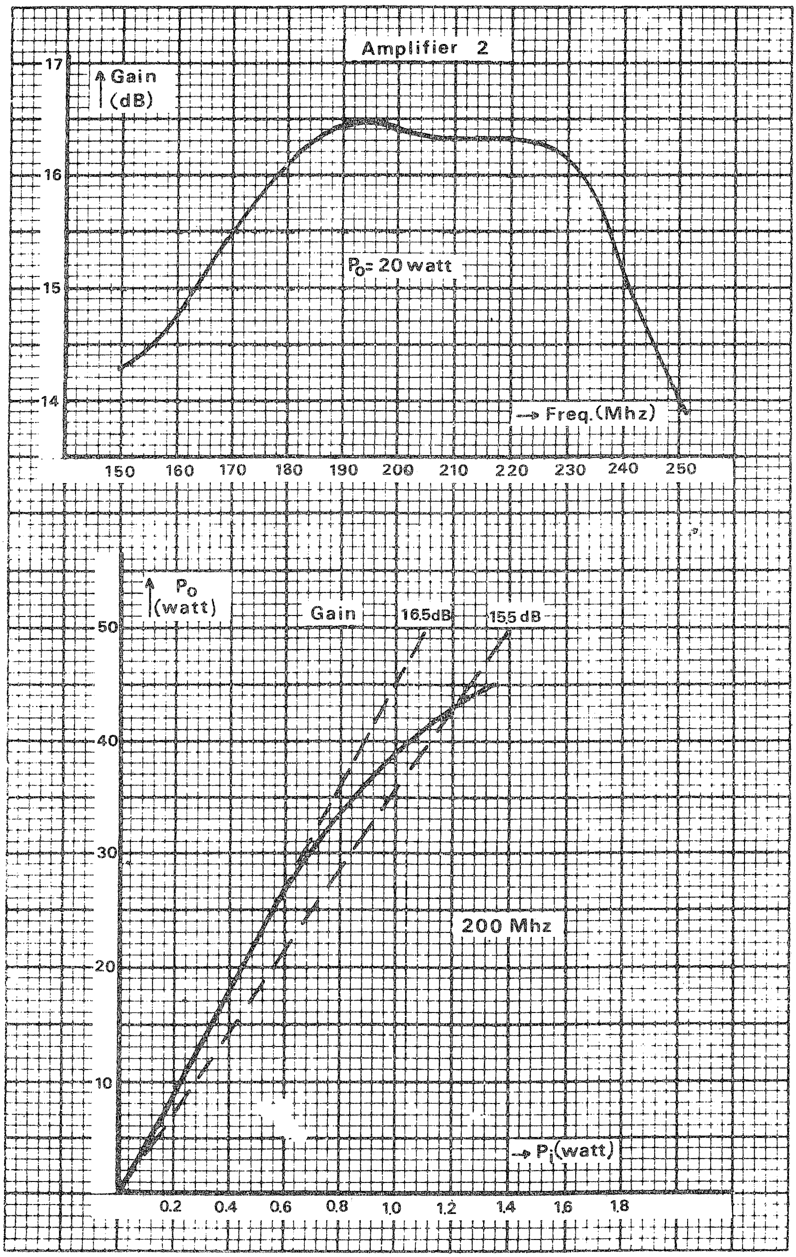

Fig.11 Gain and gain compression.

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REPORT No: NCO 8101	AUTHOR: G. Lukkassen
PROJECT No:	DATE: 26.02.1981

TITLE

A wideband hybrid coupled amplifier (470-860MHz) with two balanced transistors BLV 57

ABSTRACT

For application in TV transposers in band 4/5 (470-860MHz) a wideband linear power amplifier has been designed with two balanced transistors BLV 57 coupled by means of 3dB-90^{o} hybrids.

The class-A DC-setting of the transistors is I_C = 2x850mA and V_{CE} = 25V.

The main properties of the two prototypes are:

470-860MHz		amplifier 2	amplifier 3
gain (P_i = 1mW)	dB	9.3±0.5	9±0.5
return losses input	dB	⩾ 17	⩾ 16
return losses output	dB	⩾ 19	⩾ 19
P_o at 1dB gain compression	W	⩾ 28	⩾ 28
P_o at -55dB intermod. (3-tone, -7, -8, -16dB)	W	⩾ 17.5	⩾ 17.5
cross modulation at P_o = 15 Watt	%	⩽ 9.4	⩽ 8.8

appr. J. Tuil

Advice Patents Dept. d.d: 27-04-1981	X AV	GV		B		BL
Decision MAMO d.d: 14-04-1981	AV	X GV	X EI	B		BL
DATE: 14 APR. 1981	MAMO:					

N.V. PHILIPS SEMICONDUCTORS APPLICATION LABORATORY
NIJMEGEN - THE NETHERLANDS

REPORT No: NCO 8101	AUTHOR: G. Lukkassen
PROJECT No:	DATE: 26.02.1981

TITLE

A wideband hybrid coupled amplifier (470-860MHz) with two balanced transistors BLV 57

SUMMARY

For application in TV transposers for band 4/5 (470-860MHz) a wideband linear power amplifier has been designed with two balanced transistors BLV 57. In this hybrid coupled amplifier the BLV 57 operates in class-A with a DC-setting of I_C = 2x850mA and V_{CE} = 25V.
The circuit has been designed according to ref. 1 and ref. 2 and after that submitted to a computer optimization program.
Fig. A gives the schematic line-up of the complete amplifier.
The applied circuit board is a double copper clad PTFE fibre-glass print with an ε_r = 2.74 and a thickness of 1/32 inch.
Fig. B gives the small signal gain and return losses of the two prototypes and fig. C the gain compression.
The output power ($P_{o\ sync}$) at three intermodulation levels is given in fig. D and the cross modulation at two power levels in fig. E.
The heatsink has forced air cooling and its temperature remains below 40°C at an ambient temperature of 25°C.
The supply voltage is 28V.

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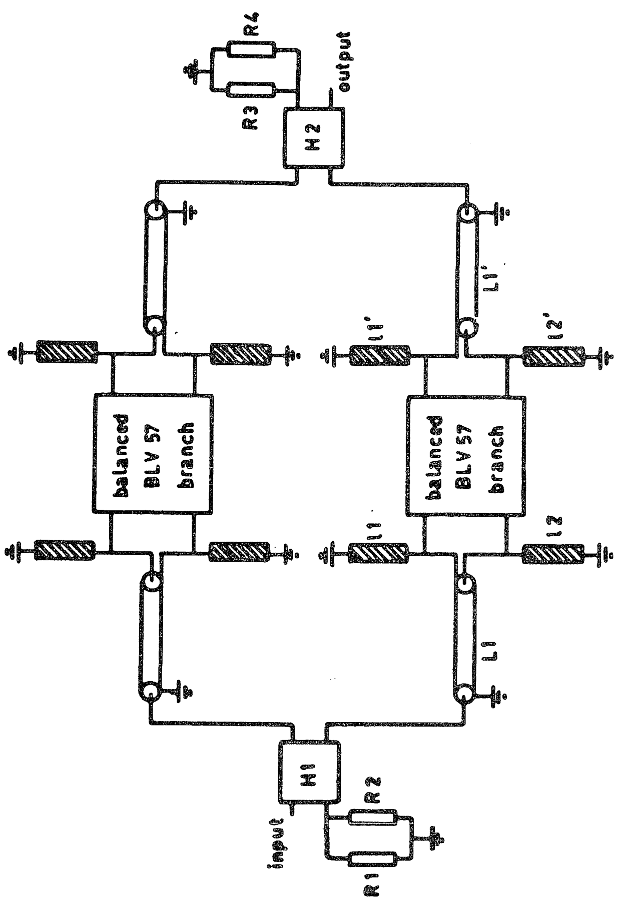
	Decision MAMO d.d:	AV	GV	EI	B		BL
	DATE: 14 APR. 1981	MAMO:					

Ref. 1: G. Matthaei
Tables of Chebychev impedance transforming networks of low-pass filter form.
Proc. of the IEEE, August 1964.

Ref. 2: O. Pitzalis Jr. and R.A. Gibson
Tables of impedance matching networks which approximate prescribed attenuation versus frequency slopes.
IEEE transactions on microwave theory and techniques, vol. MTT 119, no. 4, April 1971.



$R_1 - R_4 = 100\Omega$, CR 37 type, cat. no. 2322 212 13101

$H_1 = H_2$ = anaren hybrid coupler type: 10264-3

Fig. A: schematic line-up BLV 57 amplifier

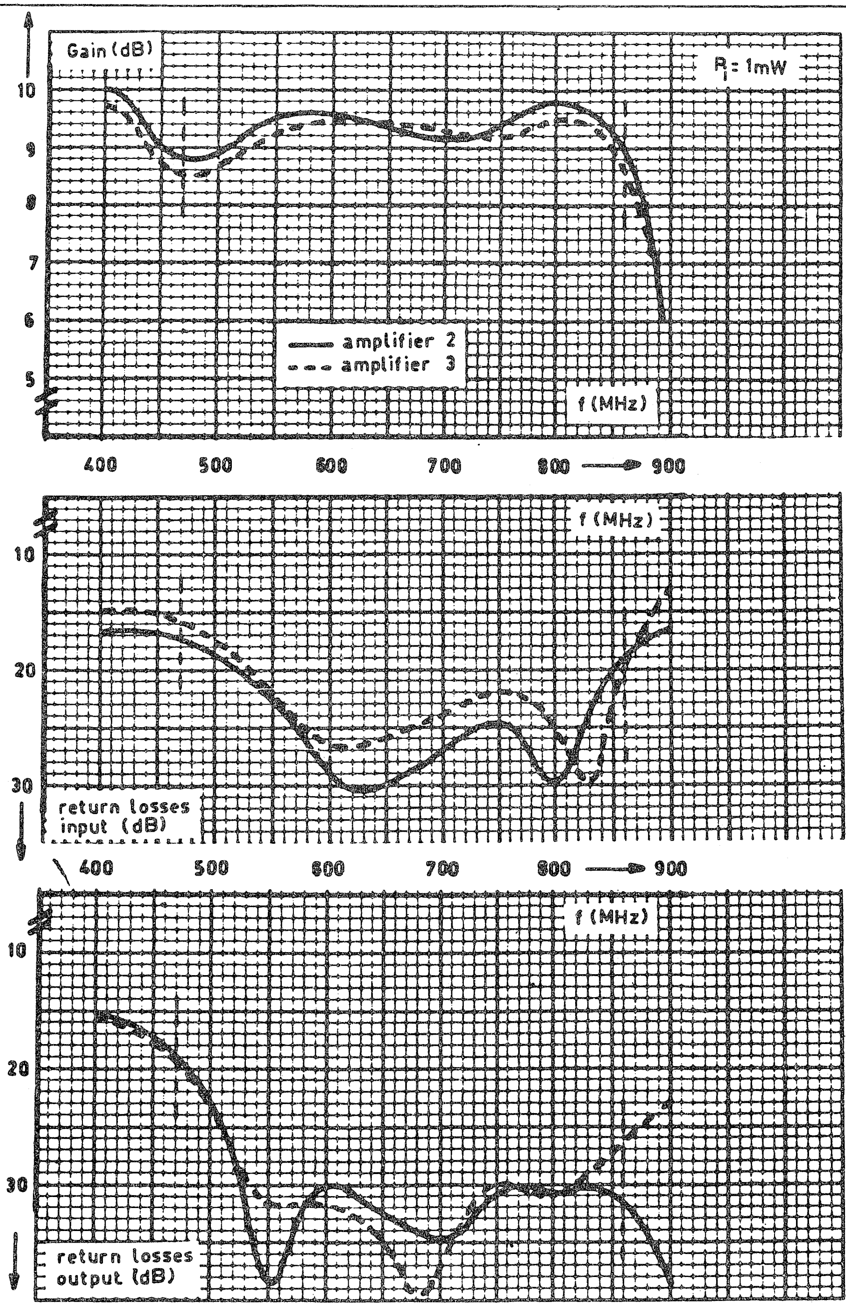
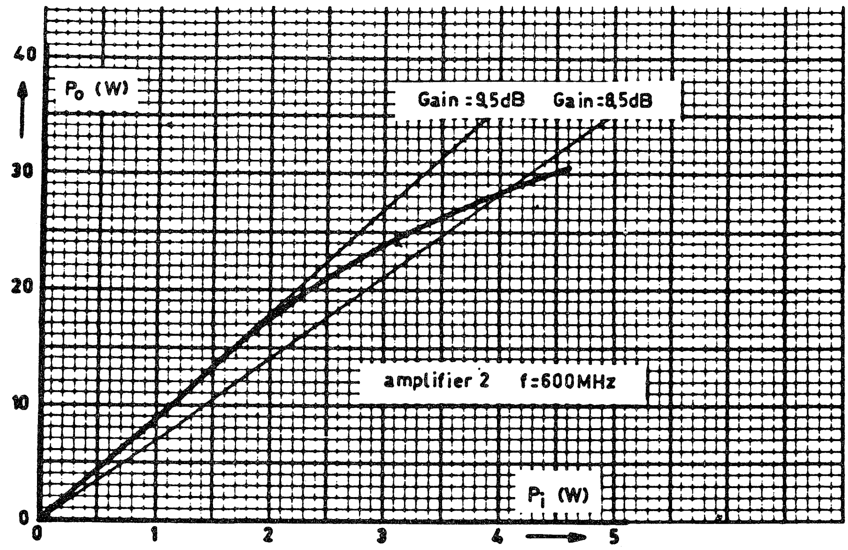


Fig. B: small signal gain and return losses



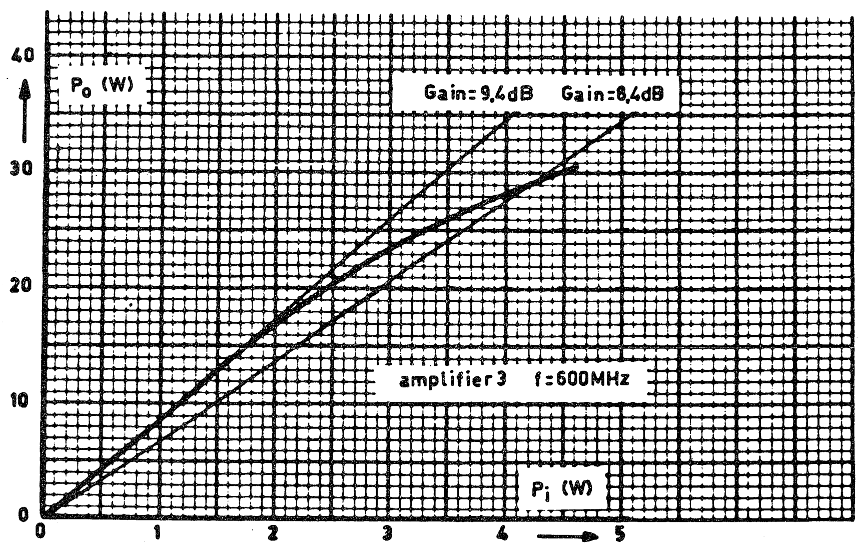


Fig. C: gain compression

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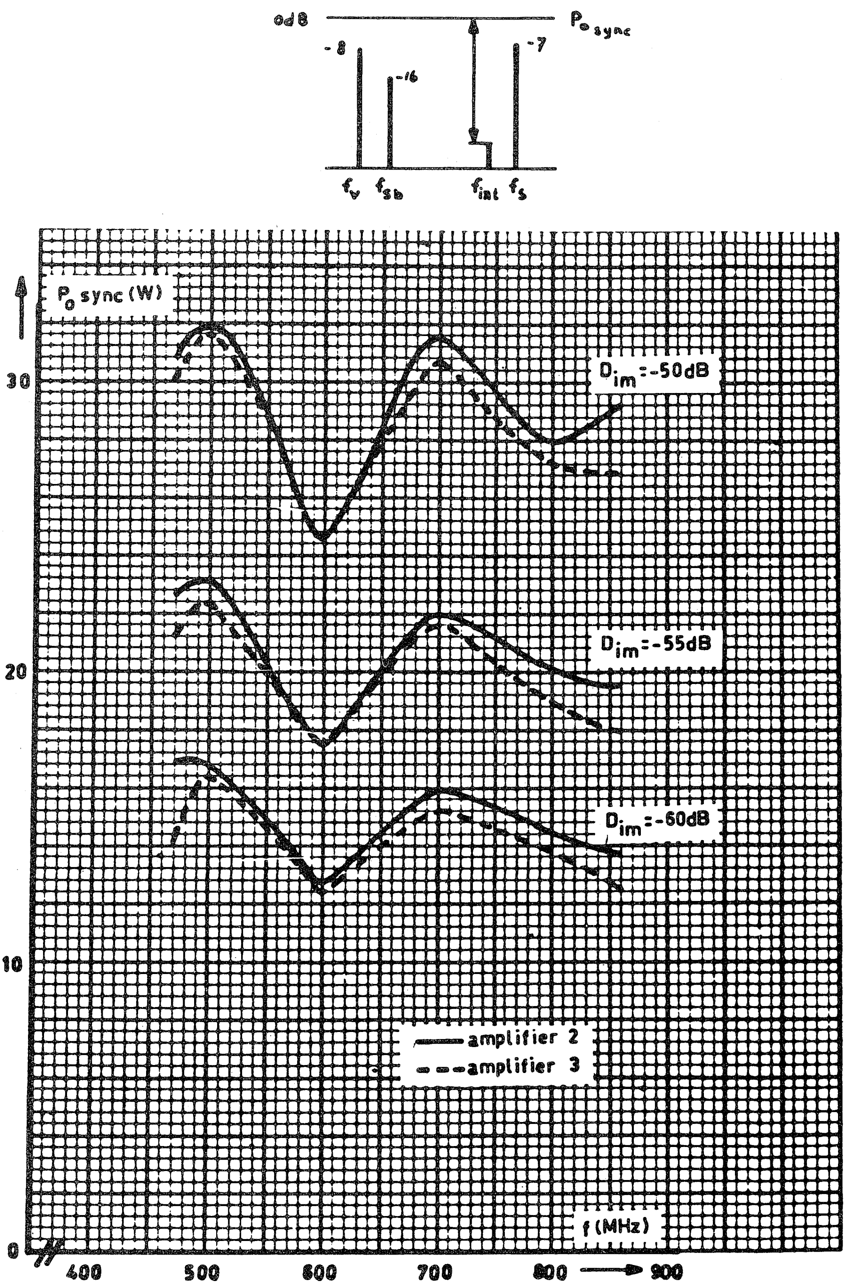

Fig. D: output power $P_{o\ sync}$

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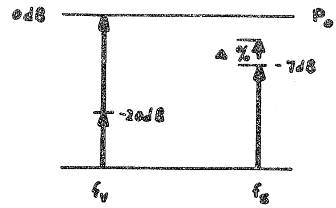

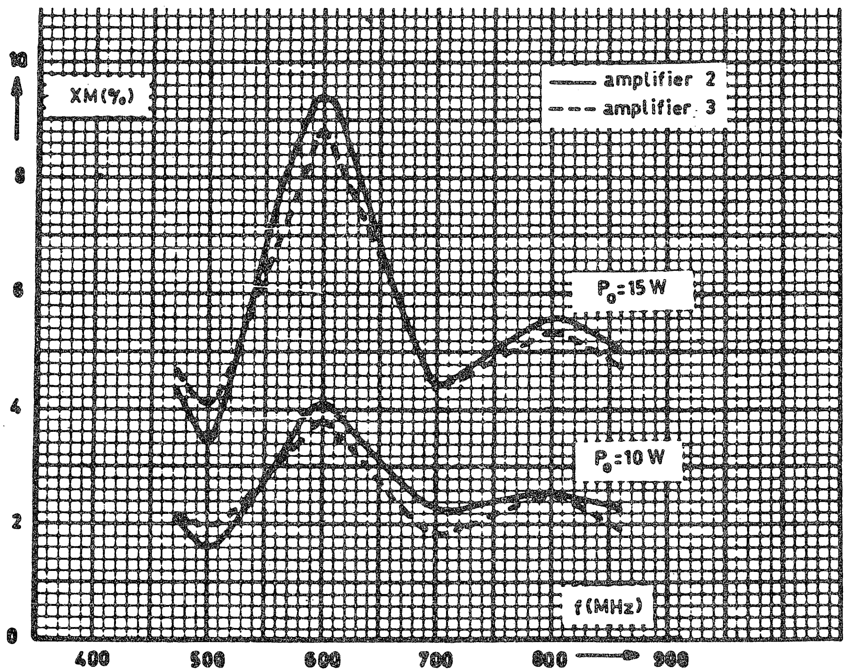

Fig. E: cross modulation

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Electronic components and materials

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Electronic components and materials

1. Introduction

For application in TV transposers for band 4/5 (470-860MHz) a wideband linear power amplifier has been designed with two transistors BLV 57 in class-A.

The BLV 57 is a balanced transistor (two identical chips) in a single envelope with a ceramic cap (NO 229).

2. Design of the amplifier

2.1 General remarks

The schematic line-up of the complete amplifier is given in fig. 1.

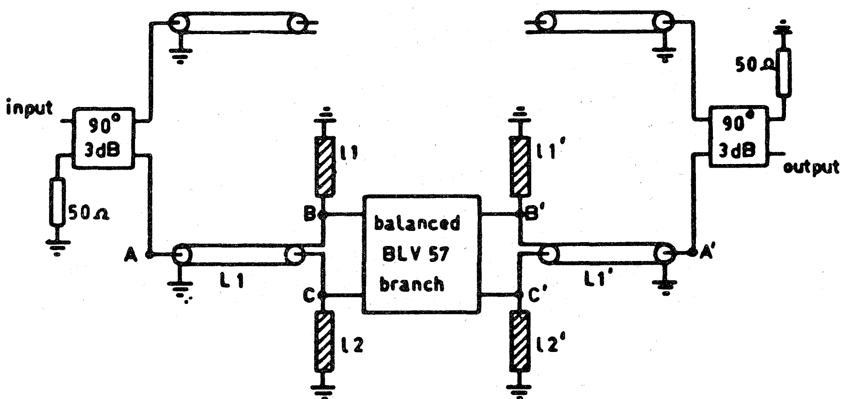

Fig. 1

The amplifier consists of 2 balanced circuits, both equipped with a BLV 57 and coupled in parallel by means of a wideband 3dB-90° coaxial hybrid at the input and output.

Each BLV 57 has 2 input circuits and 2 output circuits (one for each chip) connected to a coax balun (L_1 and L_1') which splits a 50Ω unbalanced port (A) in two 25Ω ports (B and C). The phase-shift between B and C is 180°.

The baluns (L_1 and L_1') are 50Ω semi-rigid coax cables, soldered over the whole length atop a transmission line (l_1 or l_1') of 2mm width.

To maintain circuit symmetry another shorted stub (l_2 or l'_2) with the same length has been added.
For the amplifier a PC-board has been applied of PTFE fibre-glass with an ε_r = 2.74, copper clad on both sides with a thickness of 1/32 inch.
To get a good contact between upper and lower side, rivets have been used at several places and copper straps have been soldered at the edges of the board.

2.2 Bias circuit

Each transistor has its own bias unit to obtain a stable DC-setting (see fig. 2 page 11). This bias unit enables the adjustment of the base current of each chip of the BLV 57 by means of potentiometer R_8, to obtain equal collector currents. The potentiometer R_1 adjusts both base currents simultaneous. After an accurate measurement of the values of resistors R_{12} and R_{13} the collector currents of the BLV 57, and the difference between them, can be determined easily by measuring the voltage-drop over these resistors.
The supply voltage of this bias unit is 28V.
Fig. 3 on page 12 shows the positive copy of the PC-board and the lay-out of the bias unit.

2.3 Some properties of the BLV 57

For class-A operation the BLV 57 is specified at I_C = 2x850mA and V_{CE} = 25V.
The typical gain, input and load impedance of a half BLV 57 (one chip) are given in table 1 on the next page.

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frequency (MHz)	gain (dB)	input impedance (Ω)	load impedance (Ω)
470	13.13	1.11 + j 2.62	9.75 + j 6.52
507	12.57	1.13 + j 2.88	9.00 + j 6.29
547	12.02	1.14 + j 3.17	8.26 + j 5.99
590	11.49	1.16 + j 3.47	7.52 + j 5.60
636	10.98	1.18 + j 3.80	6.80 + j 5.14
686	10.49	1.21 + j 4.16	6.10 + j 4.60
739	10.03	1.25 + j 4.55	5.44 + j 3.99
797	9.60	1.30 + j 5.00	4.80 + j 3.29
860	9.20	1.37 + j 5.50	4.20 + j 2.52

table 1

2.4 Output network

The 25Ω of the balun has to be transformed into the optimum load for the half transistor (one chip), which is given in table 1. This is done by means of an L-C output network. The circuit has been calculated according to ref. 1 and submitted to a computer optimization program.
Because the BLV 57 is a balanced transistor with two identical chips, there are two identical output circuits with a virtual ground between them.
Fig. 4 on the next page shows the calculated output circuit connected to the balun, described in section 2.1.
Two capacitors of 2.2pF and a resistor of 12Ω prevent oscillation at higher frequencies.

Electronic components and materials

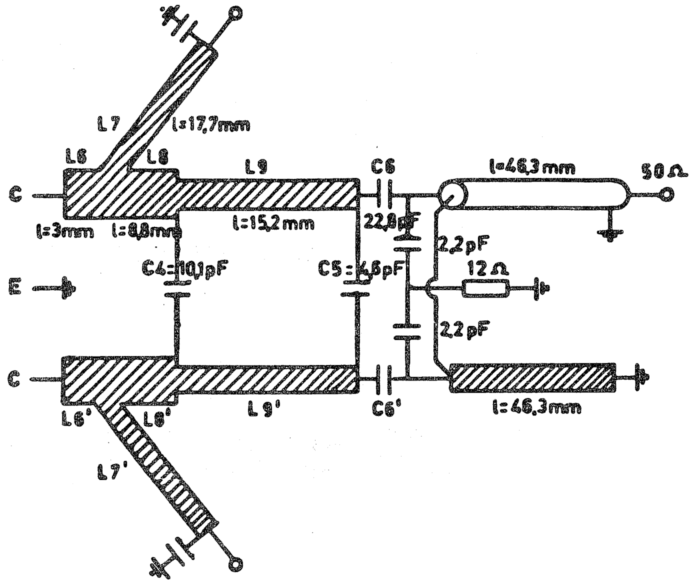


Fig. 4

The stripline L_6 (width 3mm) is the soldering place for the collector lead. The transistor is biased through the stripline L_7 (width 2mm). This stripline has been connected to L_6 and L_8 at a distance of 3mm from the transistor. L_8 is a stripline with a width of 3mm and L_9 with a width of 1.5mm.
C_4 consists of a chip capacitor in parallel with a film dielectric trimmer. C_5 is a film dielectric trimmer and C_6 is a chip capacitor.

2.5 Input network

In the frequency range of 470-860MHz the gain of a half BLV 57 (one chip) varies about 4dB (see table 1).
To decrease this gain slope an appropriate mismatch at the lower frequencies is necessary. The increasing insertion losses of the network approximately compensate the increasing gain at the lower frequencies. This method is described in ref. 2.

For the same reason as described in section 2.4 there are two identical input circuits with a virtual ground between them. Fig. 5 shows the calculated input circuit after computer optimization connected to the coaxial balun, described in section 2.1.

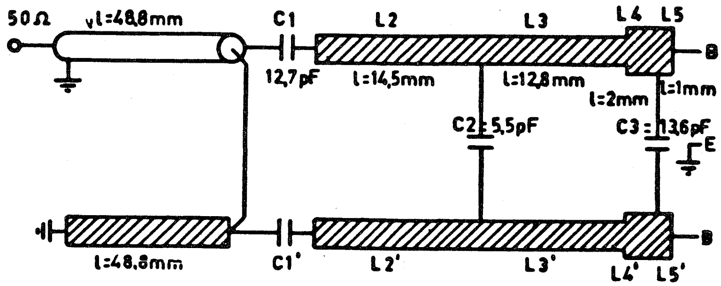


Fig. 5

The striplines L_4 and L_5 (width 3mm) form together the soldering place of the base lead.
The width of the striplines L_2 and L_3 is 1.5mm.
The capacitors C_1, C_2 and C_3 are chip capacitors.
The circuit of a complete branch (see fig. 1) is given in fig. 6 on page 13.

3. Adjustment of the amplifier

3.1 Output circuit

To obtain the highest possible output power it is essential that the transistor is given that load admittance which gives the least distortion in the used frequency range.
Therefore, for tuning, the transistor has to be replaced by a dummy consisting of a resistor and a capacitor in parallel to represent the complex conjugate of the optimum load admittance. The value of this dummy (soldered between the connection points of both collectors) has been calculated on 39Ω//8.2pF.

With the help of this dummy the output circuit has to be adjusted for good return losses at the output of L_{10} by tuning the capacitors C_{10} and C_{17} and by shifting the capacitors C_{13} and C_{14} on L_7 and L_{17} (see fig. 6 on page 13). The position of the capacitors C_{13} and C_{14} on the striplines determines the value of L_7 and L_{17}. A typical curve of the return losses at the output of L_{10} after tuning is given in fig. 7 on page 15.

3.2 Input circuit

Before tuning the input circuit, the dummy has to be replaced by the transistor, with a DC-adjustment of I_C = 2x850mA and V_{CE} = 25V. To decrease the gain-slope of the branch the input circuit has a mismatch at lower frequencies (see section 2.5).

To achieve a sufficient flat gain the capacitance of C_1, C_2, C_3 and C_4 and also the position of C_3 and C_4 (see fig. 6) can be optimized in a sweep set-up under small signal conditions (P_i = 1mW).

A typical gain curve of a balanced branch (one BLV 57) and the corresponding return losses are given in fig. 8 on page 16. It is obvious that the worse return losses at the lower frequencies are produced by the mismatch of the input circuit.

4. The hybrid coupled amplifier

In the previous section the adjustment of a balanced branch with one BLV 57 has been discussed. The gain is made flat at the cost of impermissably high return losses.

Therefore, in practice, two wideband branches are coupled in parallel by means of two wideband 3dB-90° coaxial hybrids. The properties of these hybrids reduce the return losses to at least 16dB.

The reflected power of the balanced branches is absorbed by a 50Ω resistance at the isolated port (see fig. 1).

Fig. 9 on page 17 gives a positive copy of the PC-board of the complete amplifier.

Fig. 10 on page 18 shows the lay-out of this complete amplifier.

5. Measured performance

5.1 Small signal gain and return losses

Fig. 11 on page 19 shows the gain and return losses as a function of the frequency, measured under small signal conditions. Results:

470-860MHz		amplifier 2		amplifier 3	
		min.	max.	min.	max.
gain	dB	8.8	9.8	8.5	9.5
return losses input	dB	17	30.5	16	30
return losses output	dB	17	39	18	40

table 2

5.2 Gain compression

Fig. 12 on page 20 shows the measured P_o versus P_i curves at 600MHz, which is the most critical frequency in the range. Gain compression of 1dB occurs at an output power of about 28W for both amplifiers.

5.3 Intermodulation

In fig. 13 on page 21 the output power ($P_{o\ sync}$) is given as a function of the frequency at three intermodulation levels. It has been measured according the three-tone test method (vision carrier: -8dB, sound carrier: -7dB and sideband signal: -16dB).

Zero dB corresponds to the peak sync. level. The minimum output power at -60dB intermodulation amounts to 12.4W for amplifier 2 and 12.2W for amplifier 3.

5.4 Cross modulation

Fig. 14 on page 22 shows the cross modulation as a function of the frequency at two $P_{o\ sync}$ levels.

It is a two-tone measurement (vision carrier: 0dB and sound carrier: -7dB). The amplitude of the vision carrier is changed from white level (-20dB) to peak sync level (0dB). The observed change of the voltage amplitude of the sound carrier is called cross modulation. It is expressed as a percentage of the amplitude of the sound carrier (vision carrier at white level). It has been measured with a spectrum analyser operating in linear mode.
At a $P_{o\ sync}$ level of 15W the cross modulation of amplifier 2 varies from 3.4% to 9.4% and of amplifier 3 from 4.1% to 8.8%.

6. Conclusions

It is possible to build a linear power amplifier with excellent performance with two transistors BLV 57.
The main properties of the two prototypes are:

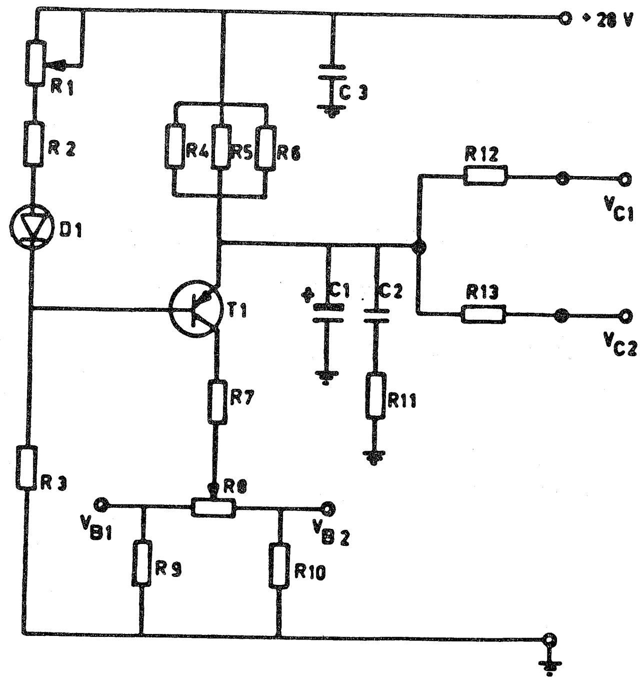
band 4/5		amplifier 2	amplifier 3
gain (P_i = 1mW)	dB	9.3±0.5	9±0.5
return losses input	dB	⩾ 17	⩾ 16
return losses output	dB	⩾ 19	⩾ 19
P_o at 1dB gain compression	W	⩾ 28	⩾ 28
$P_{o\ sync}$ at -55dB intermod. (3-tone, -7, -8, -16dB)	W	⩾ 17.5	⩾ 17.5
cross modulation at P_o=15W	%	⩽ 9.4	⩽ 8.8

table 3

7. References

Ref. 1: G.L. Matthaei
Tables of Chebychev impedance transforming networks of low pass filter form.
Proc. of the IEEE, August 1964

Ref. 2: O. Pitzalis Jr. and R.A. Gibson
Tables of impedance matching networks which approximate prescribed attenuation versus frequency slopes.
IEEE transactions on microwave theory and techniques, vol. MTT 119, no. 4, April 1971



C_1 = 10µF, 63V electrolytic capacitor, cat.no. 2222 030 28109

C_2 = 470nF, metallised film capacitor, cat.no. 4322 352 45474

C_3 = 100nF, metallised film capacitor, cat.no. 2222 352 45104

R_1 = 100Ω, cermet potentiometer, cat.no. 2122 350 00066

R_2 = 120Ω, CR 25 type, cat.no. 2322 211 13121

R_3 = 1500Ω, CR 25 type, cat.no. 2322 211 13152

R_4=R_5=R_6= 4.7Ω, enamelled wire-wound, cat.no. 2322 330 22478

R_7 = 82Ω, enamelled wire-wound, cat.no. 2322 330 22829

R_8 = 20Ω, cermet potentiometer, cat.no. 2122 350 00057

R_9=R_{10}= 39Ω, CR 25 type, cat.no. 2322 211 13399

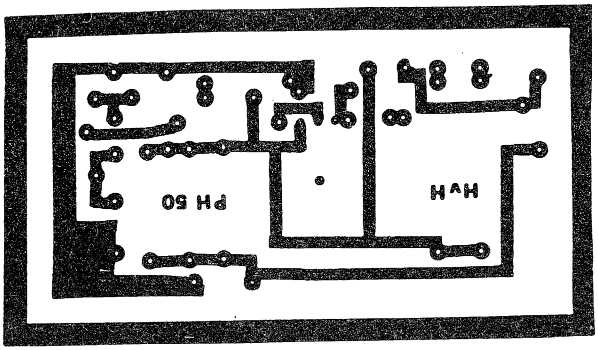
R_{11} = 10Ω, CR 25 type, cat.no. 2322 211 13109

R_{12}=R_{13}= 0.15Ω, wire-wound PM 10 type, cat.no. 2322 326 51157

D_1 = BY 206

T_1 = BD 140

Fig. 2: bias circuit



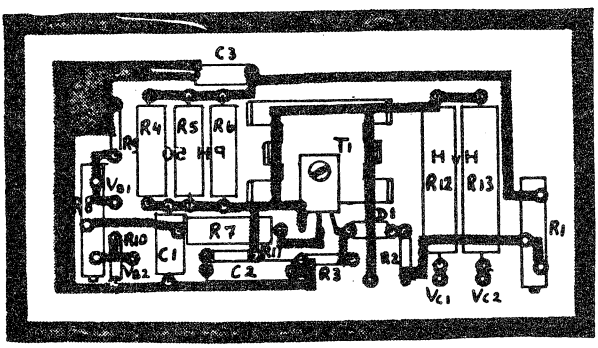


Fig. 3: lay-out and PC-board of the bias unit

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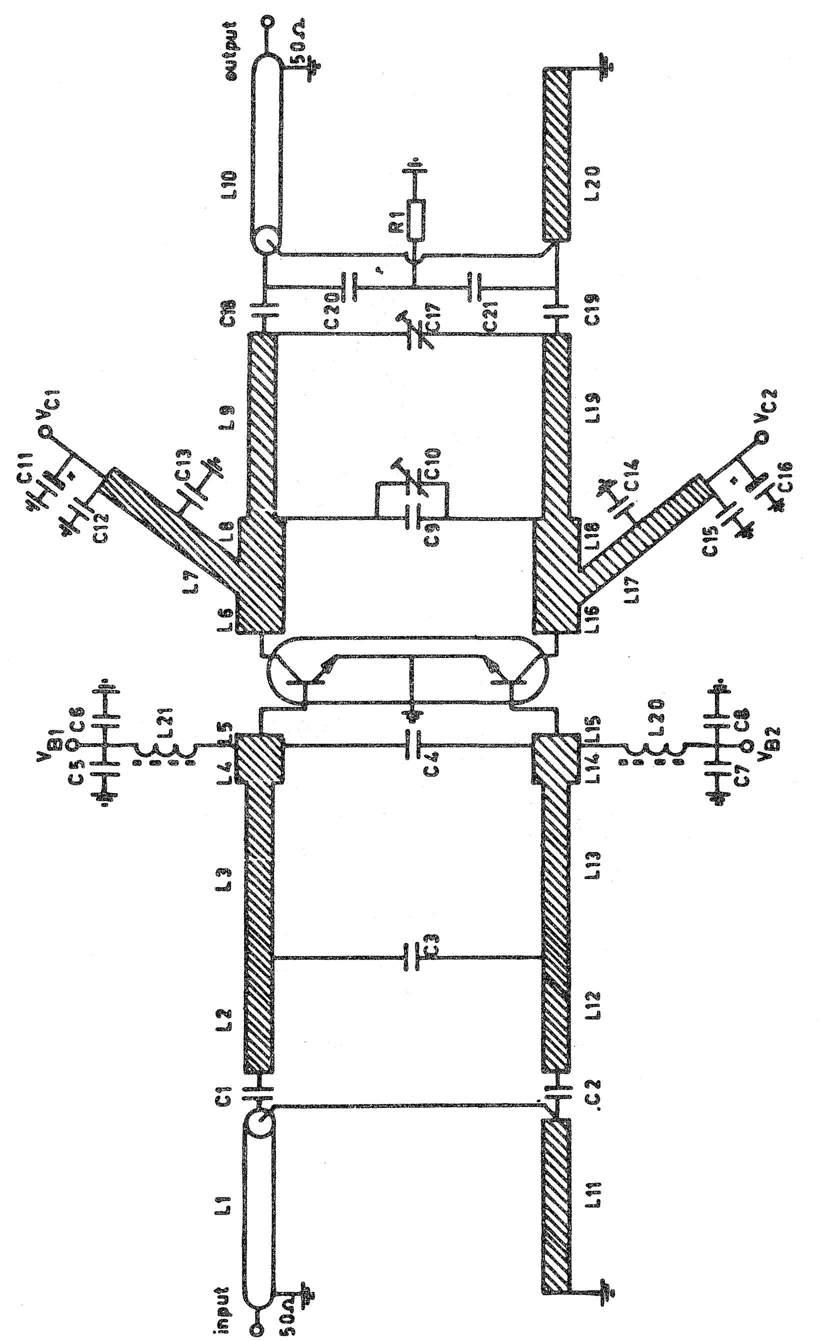

Fig. 6: circuit of a BLV 57 branch

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8. List of components BLV 57 branche

C_1=C_2=10pF chip capacitor, Philips NPO, cat.no. 2222 851 13109

C_3 = 3.9pF chip capacitor, Johanson, no. 500R 15N 3R9 CA

C_4 =12pF chip capacitor, Philips NPO, cat.no. 2222 851 13129

C_5=C_7=C_{12}=C_{15}=100nF " " , Philips NPO, cat.no. 2222 855 48104

C_6=C_8=100pF " " , Philips NPO, cat.no. 2222 852 13101

C_9 =8.2pF " " , ATC, 8R2J

C_{10}=C_{17}=1-3.5pF film dielectric trimmer, Philips, cat.no. 2222 809 05001

C_{11}=C_{16}=6.8μF, 40V, electrolytic capacitor, Philips, cat.no. 2222 030 37688

C_{13}=C_{14}=110pF chip capacitor, ATC, 111J

C_{18}=C_{19}= 22pF chip capacitor, Philips NPO, cat.no. 2222 851 13229

C_{20}=C_{21}=2.2pF chip capacitor, Johanson, no. 500R,15N 2R2 BA

L_1=49mm semi-rigid coax, 2.2mm Ø, Z_c=50Ω, PTFE dielectric soldered on 2mm stripline

L_2=L_{12}= stripline (Z_c=57Ω), 14.5x1.5mm

L_3=L_{13}= stripline (Z_c=57Ω), 12.8x1.5mm

L_4=L_{14}= stripline (Z_c=36Ω), 2 x3 mm

L_5=L_{15}= stripline (Z_c=36Ω), 1 x3 mm

L_6=L_{16}= stripline (Z_c=36Ω), 3 x3 mm

L_7=L_{17}= stripline (Z_c=48Ω), 17.7x2 mm

L_8=L_{18}= stripline (Z_c=36Ω), 8.8x3 mm

L_9=L_{19}= stripline (Z_c=57Ω), 15.2x1.5mm

L_{10}=46mm semi-rigid coax, 2.2mm Ø, Z_c=50Ω, PFTE dielectric soldered on 2mm stripline

L_{11} = stripline (Z_c=50Ω), 49x2mm

L_{20} = stripline (Z_c=50Ω), 46x2mm

L_{21}=L_{22}= 470nH micro choke, cat.no. 4322 057 04771

R=12Ω CR 25 type, cat.no. 2322 211 13129

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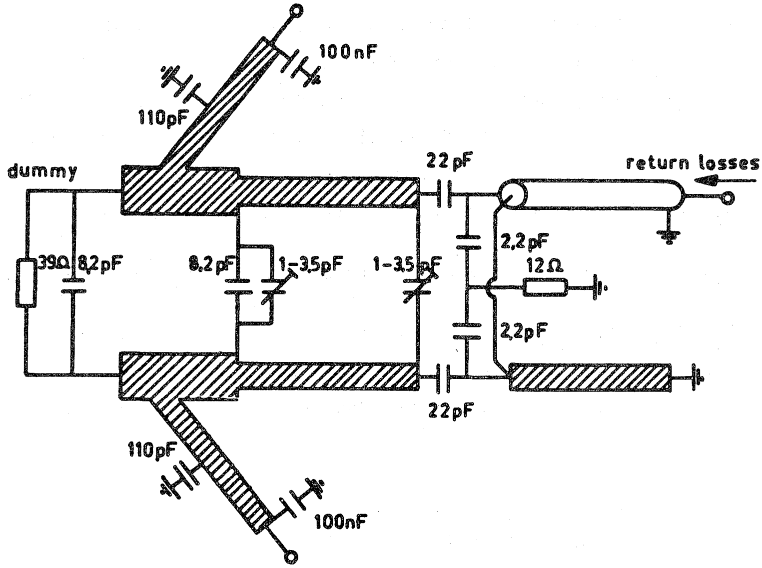

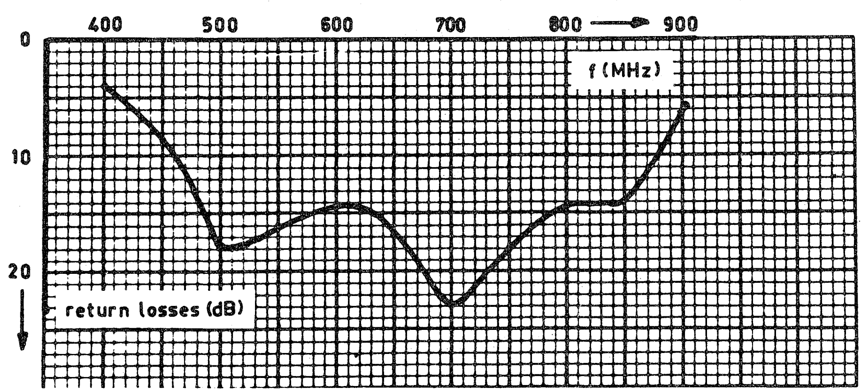

Fig. 7: tuning of the output circuit

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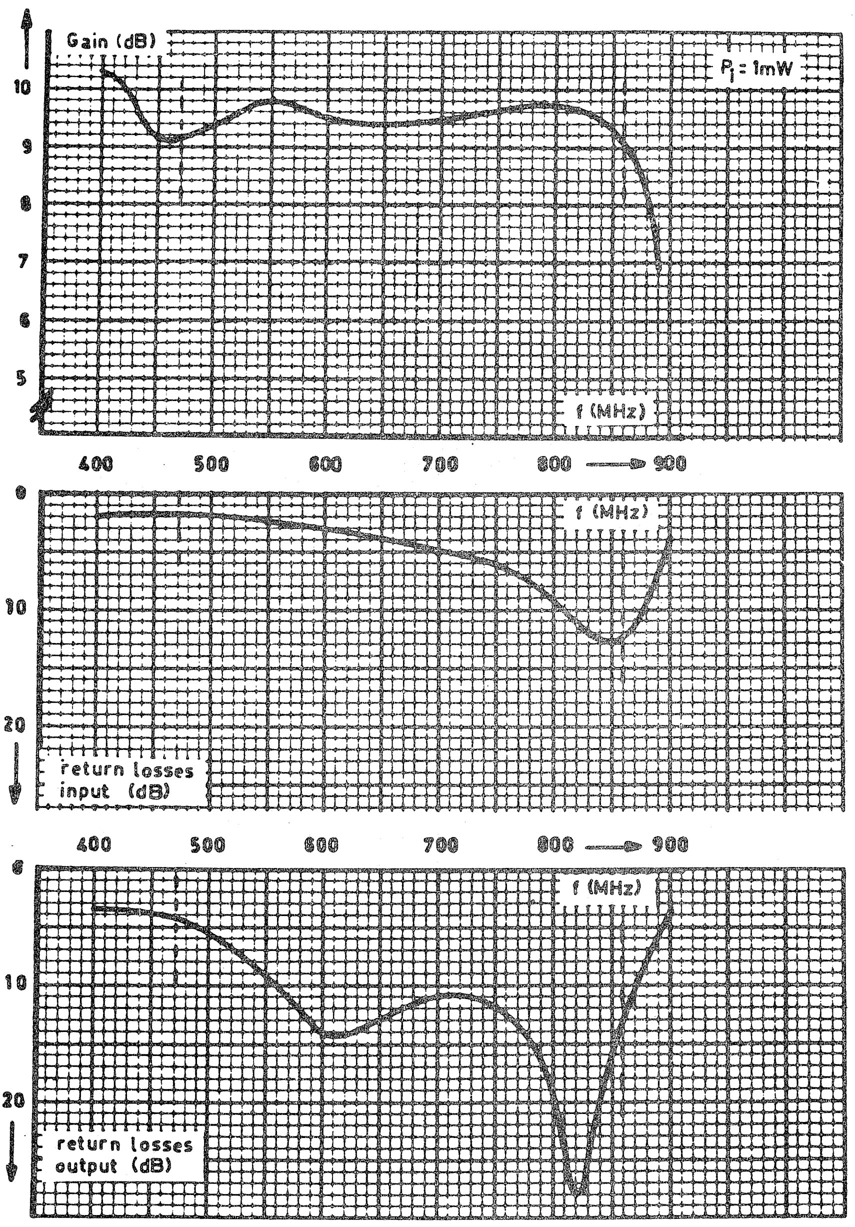

Fig. 8: gain and return losses of a BLV 57 branch

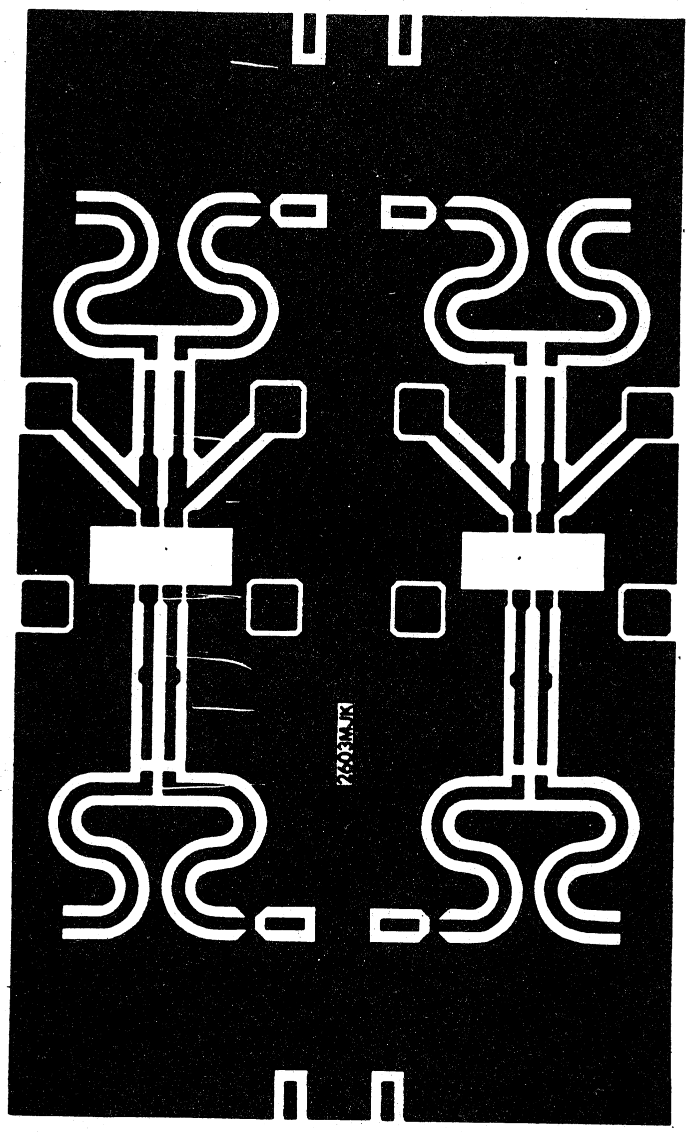

Fig. 9: PC-board of the BLV 57 amplifier

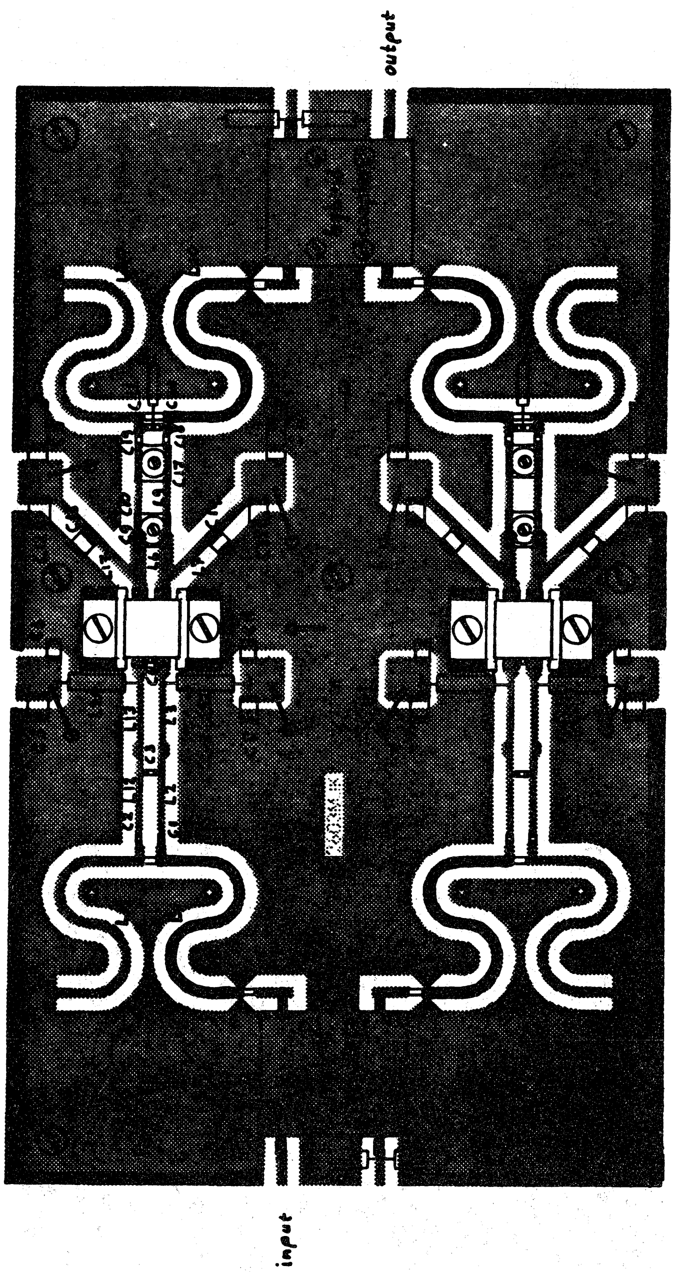

Fig. 10: lay-out of the BLV 57 amplifier

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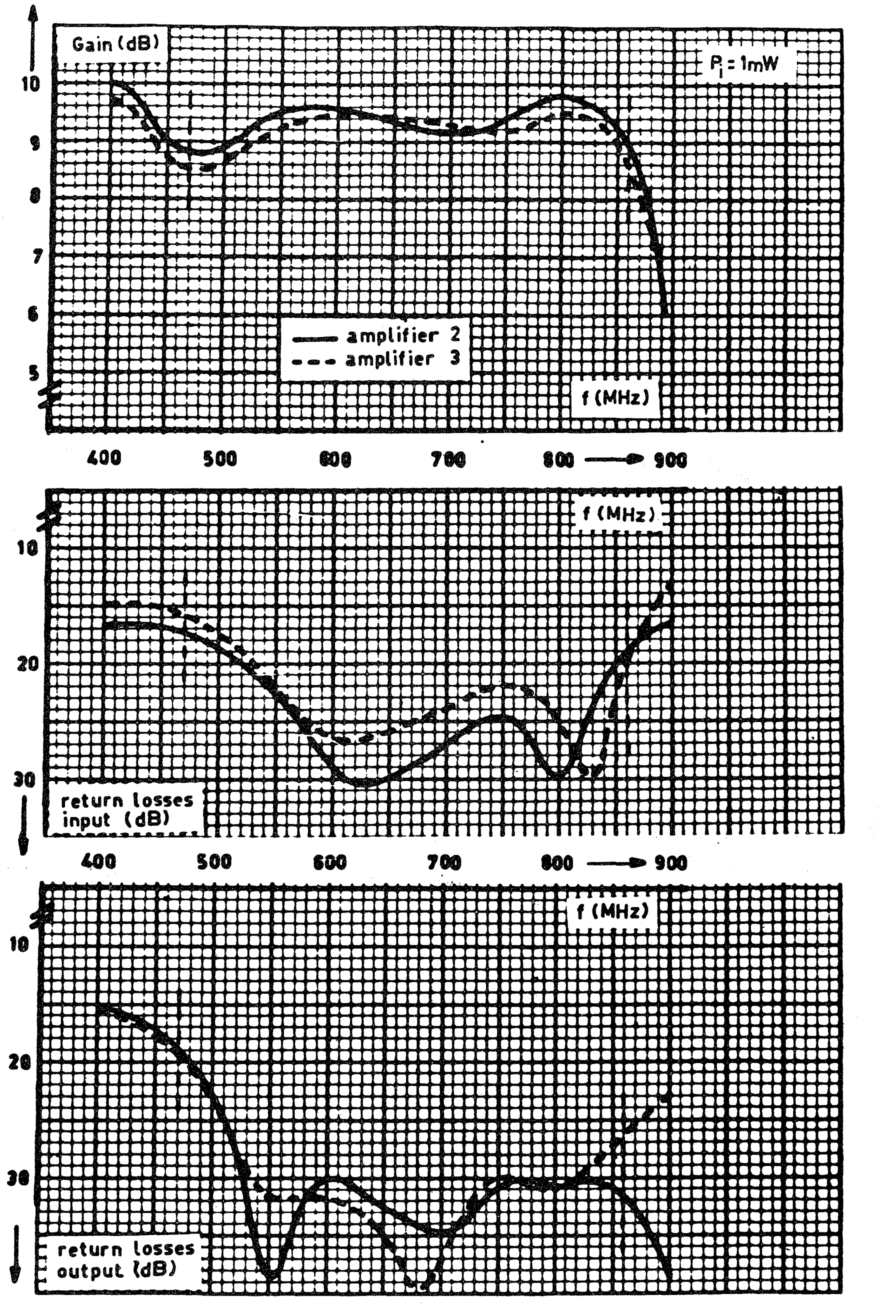
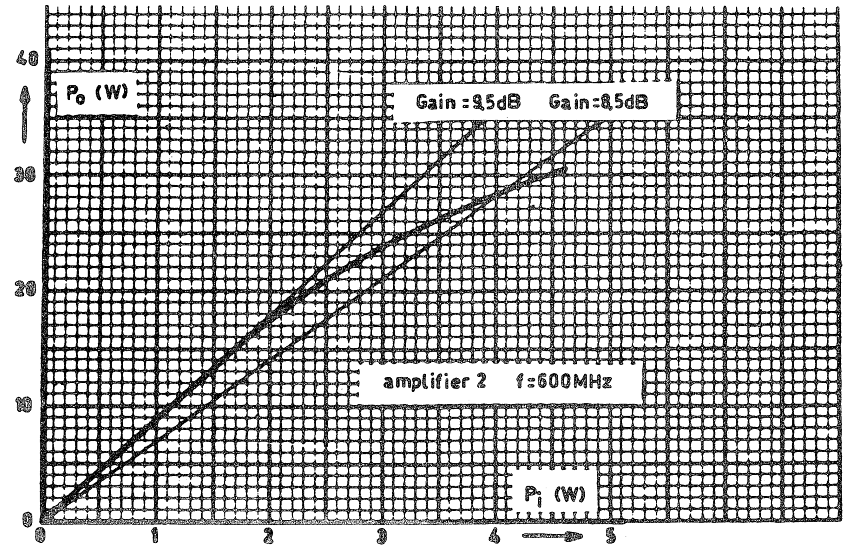


Fig. 11: gain and return losses of the BLV 57 amplifier



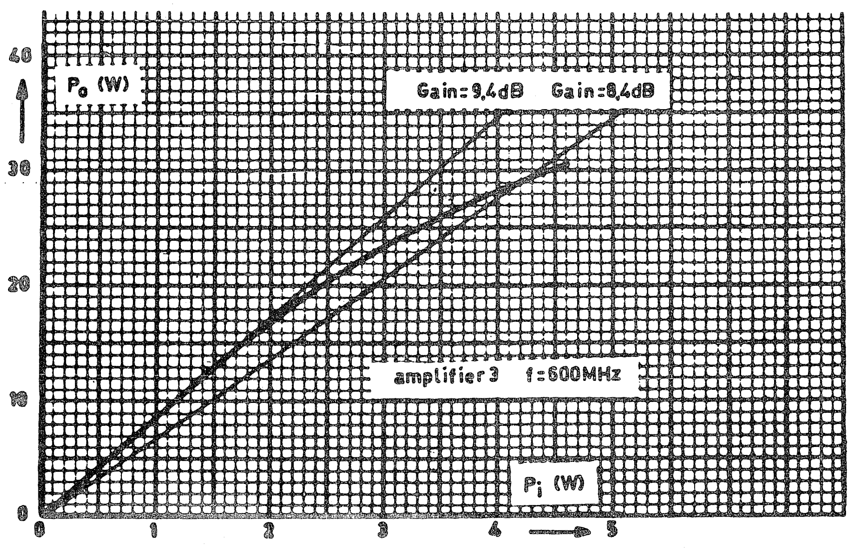


Fig. 12: gain compression of the BLV 57 amplifier

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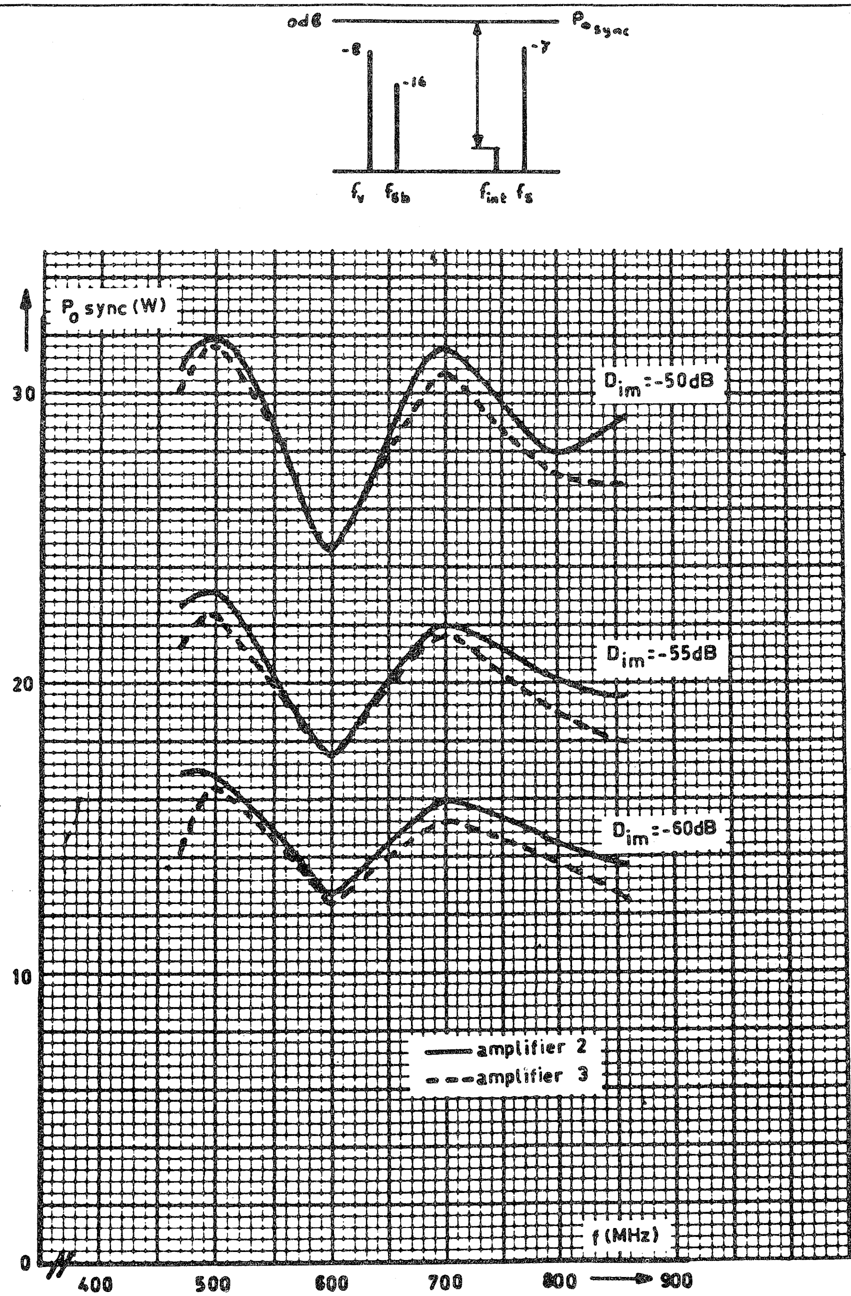

Fig. 13: output power $P_{o\ sync}$ of the BLV 57 amplifier

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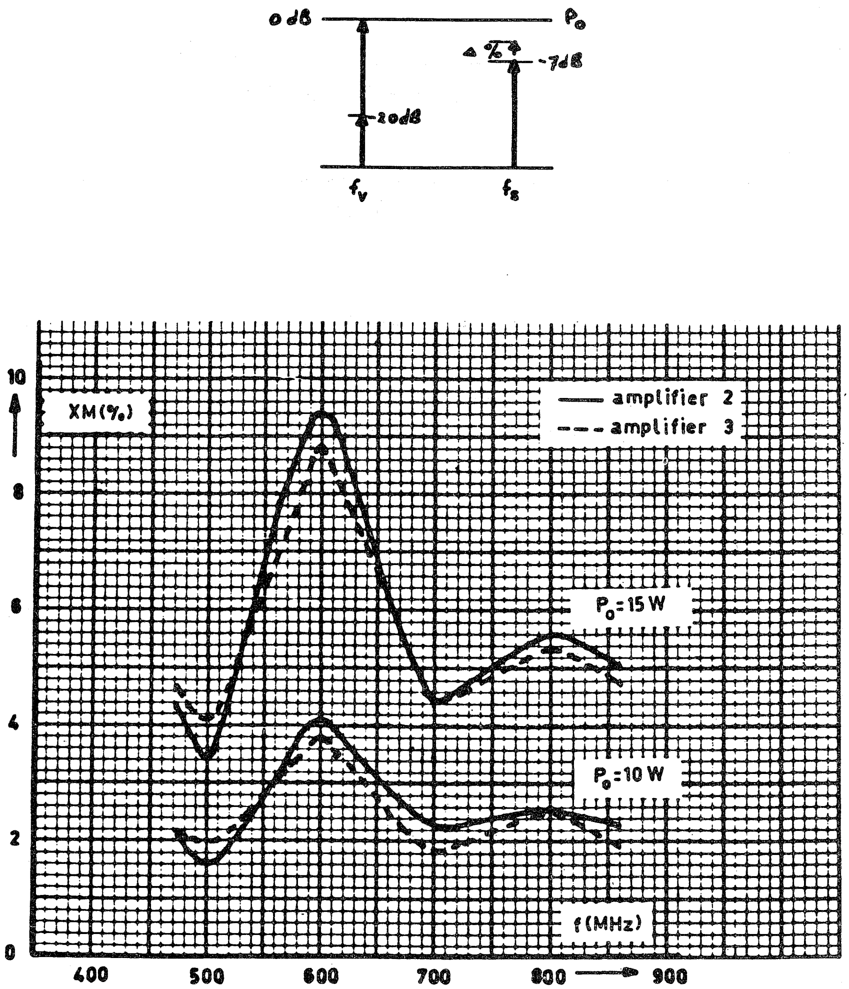

Fig. 14: cross modulation of the BLV 57 amplifier

N.V. PHILIPS SEMICONDUCTORS APPLICATION LABORATORY
NIJMEGEN – THE NETHERLANDS

REPORT No: NCO 8102	AUTHOR: G. Lukkassen
PROJECT No:	DATE: 24.11.1981

TITLE

A wideband class-A linear power amplifier (174-230MHz) with two transistors BLV 30

ABSTRACT

For application in TV transposers in band 3 (174-230MHz) a wideband amplifier has been designed with two transistors BLV 30 coupled by means of 3dB-90^{o} hybrids.

The class-A DC setting of the transistors is: I_C = 460mA and V_{CE} = 25V.

The main properties are:

gain at P_o = 2W	18.6 ± 0.3dB
return losses input	≥ 23dB
return losses output	≥ 23dB
$P_{o_{sync}}$ at -55dB intermod. (3 tone -7, -8, -16dB)	≥ 5.5W
cross modulation at P_o = 5W	≤ 9.7%
P_o at 1dB gain compression	≥ 10W

appr. J. Tuil

Advice Patents Dept. d.d: 25-01-'82	AV	GV		B		BL
Decision MAMO, d.d: 23-12-'81	AV	GV	EI	B		BL
DATE: 23 DEC. 1981	MAMO:					

N.V. PHILIPS SEMICONDUCTORS APPLICATION LABORATORY
NIJMEGEN – THE NETHERLANDS

REPORT No: NCO 8102	AUTHOR: G. Lukkassen
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TITLE

A wideband class-A linear power amplifier (174-230MHz) with two transistors BLV 30

SUMMARY

For application in driver stages of TV transposers in band 3 a wideband class-A linear power amplifier with two transistors BLV 30 has been designed.
The DC setting is: I_C = 400mA and V_{CE} = 25V.
The amplifier consists of two branches coupled with two 3dB-90^{o} coaxial hybrid couplers.
The theoretical design of the circuit is described in report ECO 8004 written by Mr. A.H. Hilbers.
The applied circuit board is a double copper clad print of epoxy fibre-glass with a thickness of 1/16 inch (ε_r = 4.5)
Practical optimisation and tuning has been done on a dynamic gain compression set-up.
The main properties are:

gain at P_o = 2W	18.6 $\pm$ 0.3dB
return losses input	$\geqslant$ 23dB
return losses output	$\geqslant$ 23dB
$P_{o_{sync}}$ at -55dB intermod. (3 tone, -7, -8, -16dB)	$\geqslant$ 5.5W
cross modulation at P_o = 5W	$\leqslant$ 9.7%

Advice Patents Dept. d.d: 05-01-82	X AV	GV		B		BL	
Decision MAMO d.d: 23-12-81	AV	X GV	X EI	B		BL	
DATE:	MAMO:						

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P_o at 1dB gain compression $\geqslant$ 10W

The heatsink temperature remains below 40^{o}C at an ambient temperature of 25^{o}C.

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1. Introduction

For application in driver stages of TV transposers in band 3 a wide-band linear amplifier has been designed with two transistors BLV 30. The theoretical design has been given in report ECO 8004 written by Mr. A.H. Hilbers.

The transistors operate in class-A at V_{CE} = 25V and I_C = 460mA.

The encapsulation is a ¼ inch stud type with a ceramic cap (SOT 122).

2. Design of the amplifier

2.1 General remarks

The amplifier consists of 2 circuits, both equipped with a BLV 30 and coupled in parallel by means of a wideband 3dB-90^o coaxial hybrid coupler at the input and the output.

For the amplifier a PC board has been applied of epoxy fibre-glass with an ε_r = 4.5, copper clad on both sides with a thickness of 1/16 inch. To get a good contact between upper and lower side, rivets have been used at several places and copper straps have been soldered at the edges of the board. To reach a small emitter inductance also straps have been soldered at the emitter contacts to the lower side of the PC board.

2.2 Bias circuit

Each transistor has its own bias unit to obtain a stable DC setting. These bias units have been mounted atop the PC board of the amplifier.

The supply voltage of these 2 bias units is 27.5V.

Fig. 1 on page 7 shows the positive copy of the PC board, the layout and the circuit of a bias unit.

2.3 Some properties of the BLV 30

For class-A operation the BLV 30 is specified at I_C = 460mA and V_{CE} = 25V.

The typical gain, input- and load impedance are given in table 1 on the next page.

table 1

frequency (MHz)	gain (dB)	input impedance (Ω)	load impedance (Ω)
174	21.3	1.08 + j0.29	20.8 + j19.6
202	20	1.07 + j0.53	17.8 + j19.5
230	18.8	1.07 + j0.74	15.2 + j19.0

2.4 Output network

The 50Ω system impedance has to be transformed into the optimum load for the transistor over the frequency range 174-230MHz. This is done by means of an L-C output network. The circuit has been calculated and submitted to a computer optimization program (see ref. 1).

Fig. 2 shows the calculated h.f. output circuit.

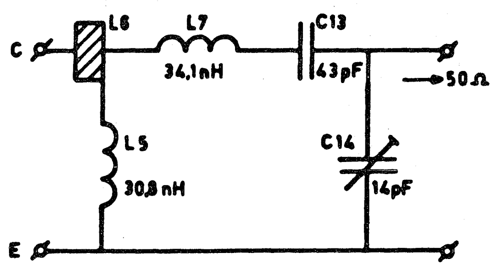


fig. 2

The small 30Ω stripline L_6 (w = 6mm, l = 3mm) is the soldering place for the collector lead.

The transistor is biased through the inductance L_5 which resonates with the transistor output capacitance in the middle of the frequency band.

2.5 Input network

In the frequency range of 174-230MHz the gain of the BLV 30 varies about 2.5dB (see table 1).

To decrease the gain slope the input circuit has a mismatch at the lower frequencies and a matching at the high end of the frequency band.
Fig. 3 shows the calculated h.f. input circuit after computer optimization.

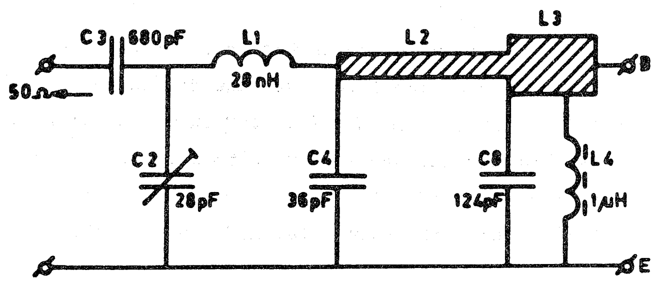


fig. 3

The inductances L_2 and L_3 are executed as striplines. The stripline L_2 = 35mm long and 2mm wide (Z_c = 60Ω). The stripline L_3 is 12.9mm long and 6mm wide (Z_c = 30Ω) and is also the soldering place for the base lead. The transistor is biased through the inductance L_4.

3. Tuning of the amplifier

Tuning of the amplifier takes place before coupling the two branches with the 3dB-90° hybrid couplers.
Each individual branch has been optimized by means of a powersweep set-up. In this case the hybrid couplers have been replaced by 2 short lengths of 50Ω semi rigid cable connecting the input to C_3 and the output to C_{13}, C_{14} and C_{15} (see fig. 4 on page 8).
In the powersweep set-up it is possible to measure gain compression and output power under swept conditions.
Because gain compression correlates with the distortion, in this powersweep set-up it is possible to find the best compromise between a sufficiently flat gain and a high power at a low distortion level.

Minimum gain compression and sufficiently flat gain have been established by trimming the capacitors C_2 and C_{14} (see fig. 4 on page 8).

4. The hybrid coupled amplifier

In the previous section the tuning of a branch with one BLV 30 has been discussed. The gain is made flat by accepting high input return losses. This problem is solved by coupling the two branches in parallel by means of two 3dB-90^o coaxial hybrid couplers. The properties of these couplers reduce the return losses to at least 23dB. The reflected power of the balanced branches is absorbed by a 50Ω resistance at the isolated port (see fig. 4 on page 8).

Fig. 5 on page 10 gives the positive copy of the PC board of the complete amplifier.

Fig. 6 on page 11 shows the layout of this complete amplifier.

5. Measured performance

5.1 Small signal gain and return losses

Fig. 7 on page 12 shows the gain and return losses as a function of the frequency, measured under small signal condition.

The gain is min. 18.3dB and max. 18.9dB.

The return losses are at least 23dB.

5.2 Intermodulation

In fig. 8 on page 13 the output power ($P_{o_{sync}}$) is given as a function of the frequency at three intermodulation levels. It has been measured according the three tone test method (vision carrier: -8dB, sound carrier: -7dB and sideband signal: -16dB). Zero dB corresponds to the peak sync. level.

The minimum output power at -55dB is 5.5 Watt.

5.3 Cross modulation

Fig. 9 on page 14 shows the cross modulation as a function of the frequency at five $P_{o_{sync}}$ levels.

It is a two tone measurement (vision carrier: 0dB and sound carrier: -7dB). The amplitude of the vision carrier is changed from white level (-20dB) to peak sync. level (0dB). The observed change of the voltage amplitude of the sound carrier is called cross modulation. It is expressed as a percentage of the amplitude of the sound carrier (vision carrier at white level) and has been measured with a spectrum analyser operating in linear mode.

At a $P_{o_{sync}}$ level of 7 Watt the cross modulation varies from about 15% to 20% and at 5 Watt from 7.9% to 9.7%.

5.4 Gain compression

Fig. 10 on page 15 shows the measured P_o versus P_i curves at 174MHz. Measurements showed this is the most critical frequency in the range.

Gain compression of 1dB occurs at an output power of 10 Watt.

6. Conclusions

The in this report described amplifier has excellent performances.

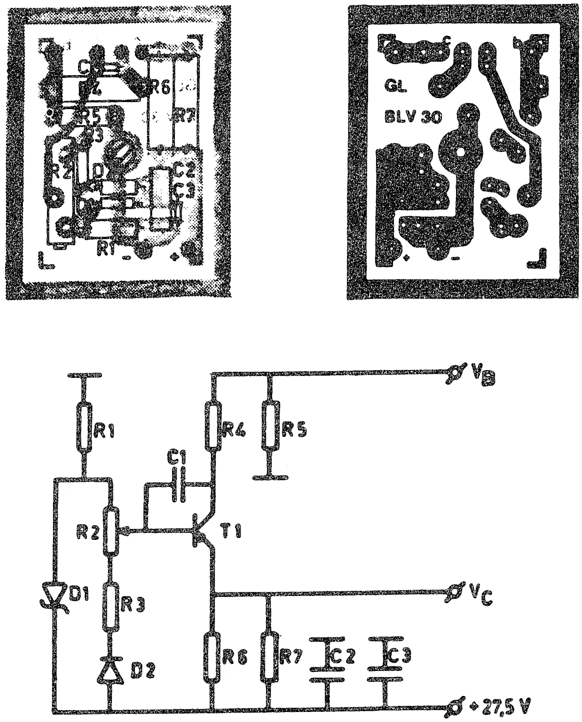
The main properties of the band 3 amplifier are:

gain (P_o = 2W)	18.6 $\pm$ 0.3dB
return losses input	$>$ 23dB
return losses output	$>$ 23dB
$P_{o_{sync}}$ at -55dB intermod. (3 tone, -7, -8, -16dB)	$>$ 5.5W
cross modulation at P_o = 5W	$<$ 9.7%
P_o at 1dB gain compression	$>$ 10W

7. References

ref. 1: G.L. Matthaei

Tables of Chebychev impedance transforming networks of low pass filter form. Proc. of the IEEE, August 1964.



R_1 = 1000Ω, CR 37 type cat.no. 2322 212 13102

R_2 = 100Ω, cermet potentiometer cat.no. 2122 350 00066

R_3 = 75Ω, CR 25 type cat.no. 2322 211 13759

R_4 = 100Ω, enamelled wire wound cat.no. 2322 330 22101

R_5 = 68Ω, CR 37 type cat.no. 2322 212 13689

R_6 = R_7 = 10Ω, enamelled wire wound cat.no. 2322 330 22109

C_1 = 100pF, ceramic capacitor cat.no. 2222 632 10101

C_2 = 100nF, metallised film cap. cat.no. 2222 352 45104

C_3 = 1000pF, ceramic capacitor cat.no. 2222 630 02102

D_1 = BZX 79, 3V6

D_2 = BY 206

T_1 = BD 140

fig. 1: BLV 30 bias circuit and layout scale 1:1

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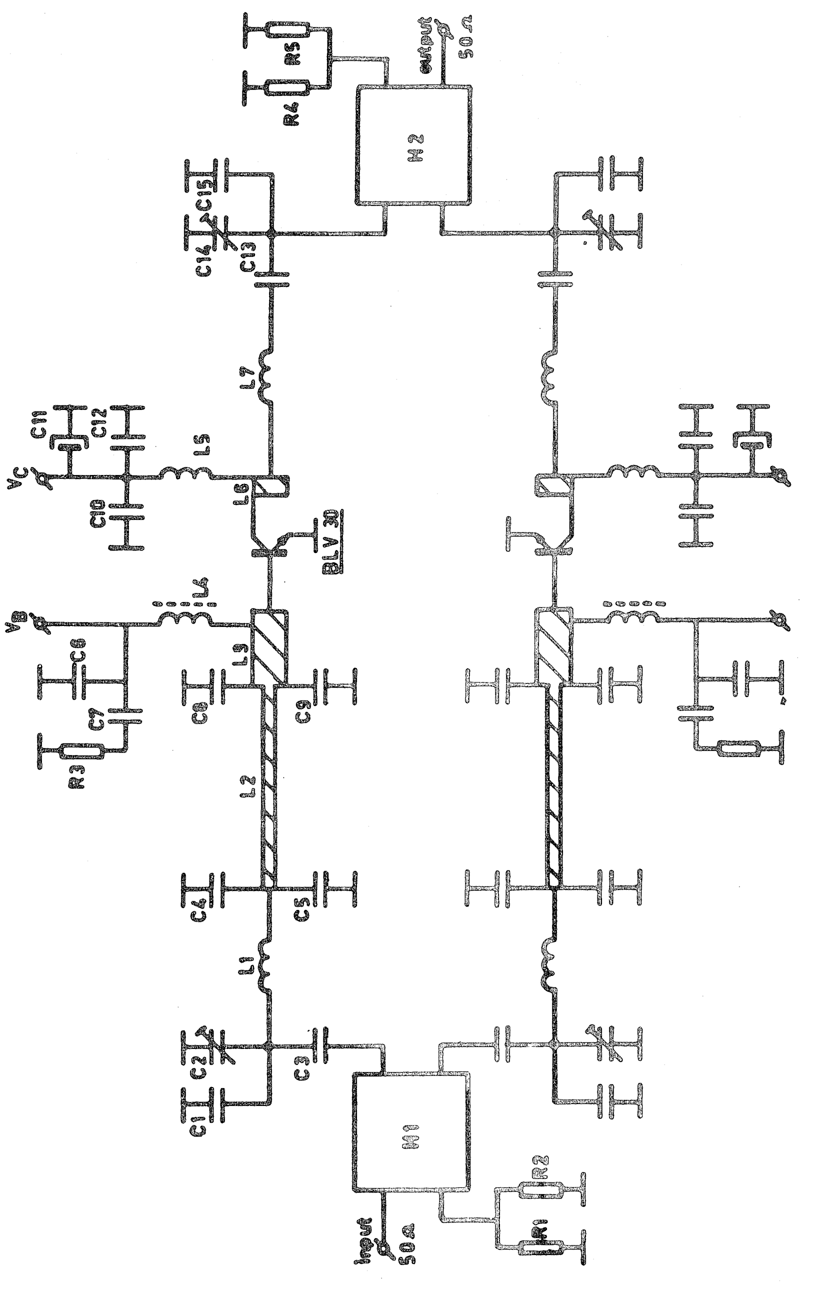

fig. 4: circuit of the BLV 30 amplifier

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8. List of components of the BLV 30 amplifier

C_1 = 10pF, chip capacitor, NPO size 0805	cat.no. 2222 851 13109
C_2 = 2-18pF, film dielectric trimmer	cat.no. 2222 809 05003
$C_3=C_6=C_{10}$=680pF, chip capacitor, NPO size 1210	cat.no. 2222 852 13681
$C_4=C_5$=18pF, chip capacitor, NPO size 0805	cat.no. 2222 851 13189
$C_7=C_{12}$=330nF, metallised film capacitor	cat.no. 2222 352 45334
C_8 = 68pF, chip capacitor, NPO size 0805	cat.no. 2222 851 13689
C_9 = 56pF, chip capacitor, NPO size 0805	cat.no. 2222 851 13569
C_{11}= 10µF, electrolytic capacitor	cat.no. 2222 030 28109
C_{13}= 43pF, ATC chip capacitor	100B-430-J-P-X-500
C_{14}= 2-9pF, film dielectric trimmer	cat.no. 2222 809 05002
C_{15}= 5.6pF, ATC chip capacitor	100A-5R6-B-P-X-50

$R_1=R_2=R_4=R_5$= 100Ω, CR 37 type	cat.no. 2322 212 13101
R_3 = 10Ω, CR 37 type	cat.no. 2322 212 13109

L_1 = 28nH, 3 turns of 1mm copper wire, int. dia. = 3mm, length = 4.1mm

L_2 = stripline (Z_c = 60Ω), 35x2mm

L_3 = stripline (Z_c = 30Ω), 12.9x6mm

L_4 = 1µH, microchoke cat.no. 4322 057 01081

L_5 = 30.8nH, 4 turns of 1mm copper wire, int.dia. = 3mm, length = 7.6mm

L_6 = stripline (Z_c = 30Ω), 3x6mm

L_7 = 34.1nH, 4 turns of 1mm copper wire, int.dia. = 3mm, length = 6.6mm

$H_1=H_2$= Anaren 3dB-90° hybrid coupler, model no. 10262-3

PC board material: epoxy fibre-glass (ε_r = 4.5), thickness: 1/16 inch

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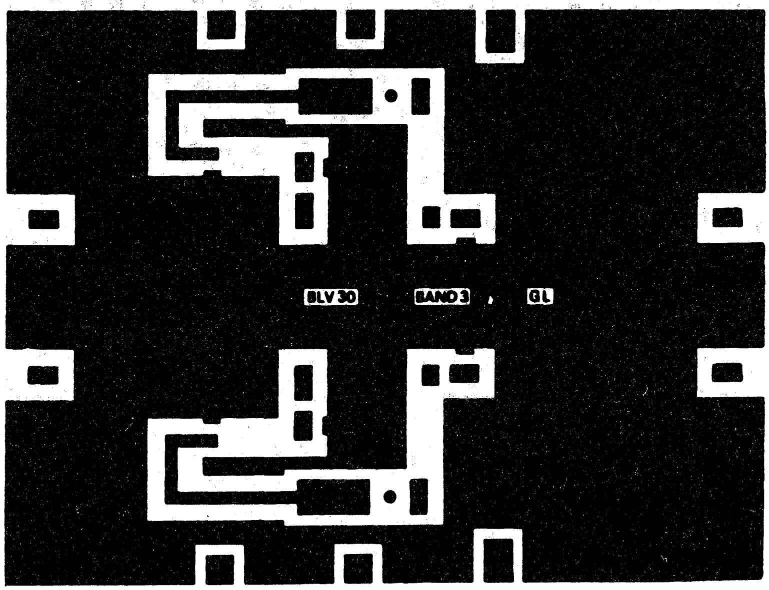

fig. 5: PC board of the BLV 30 amplifier

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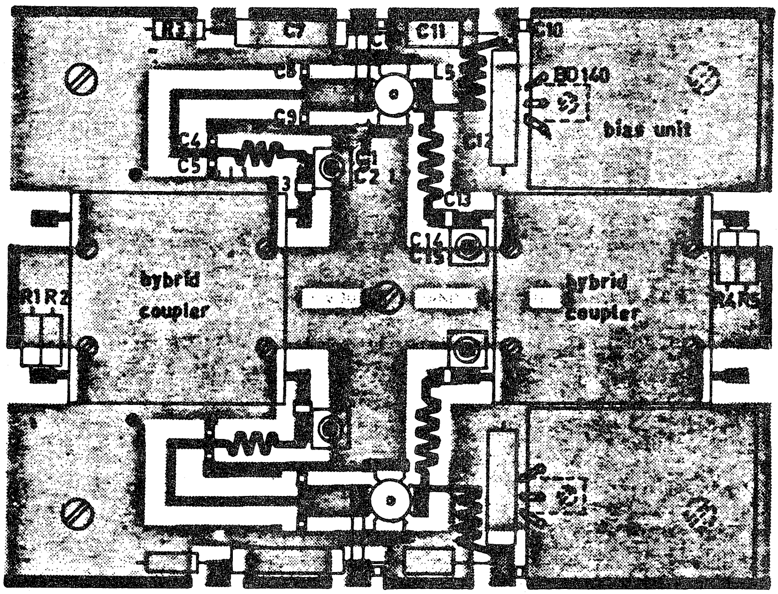

fig. 6: layout of the BLV 30 amplifier

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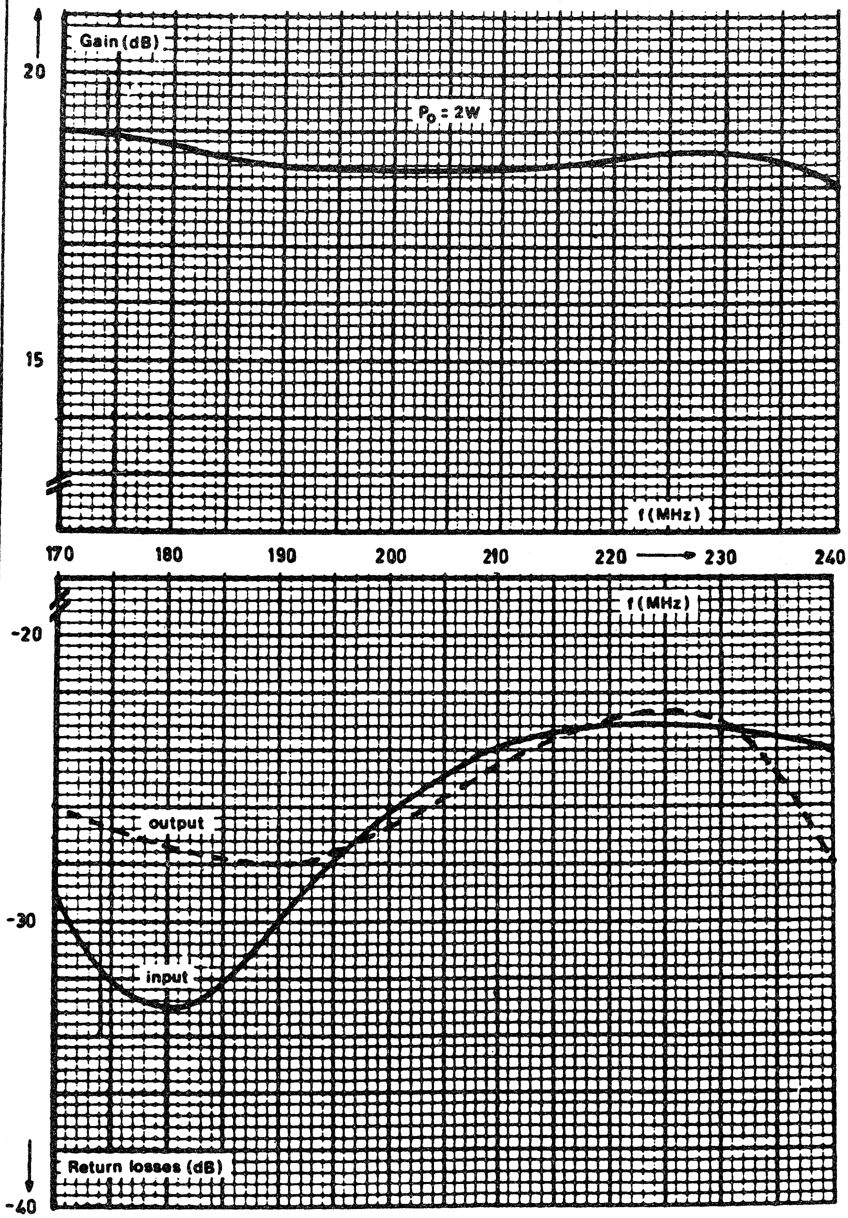

fig. 7: gain and return losses of the BLV 30 amplifier

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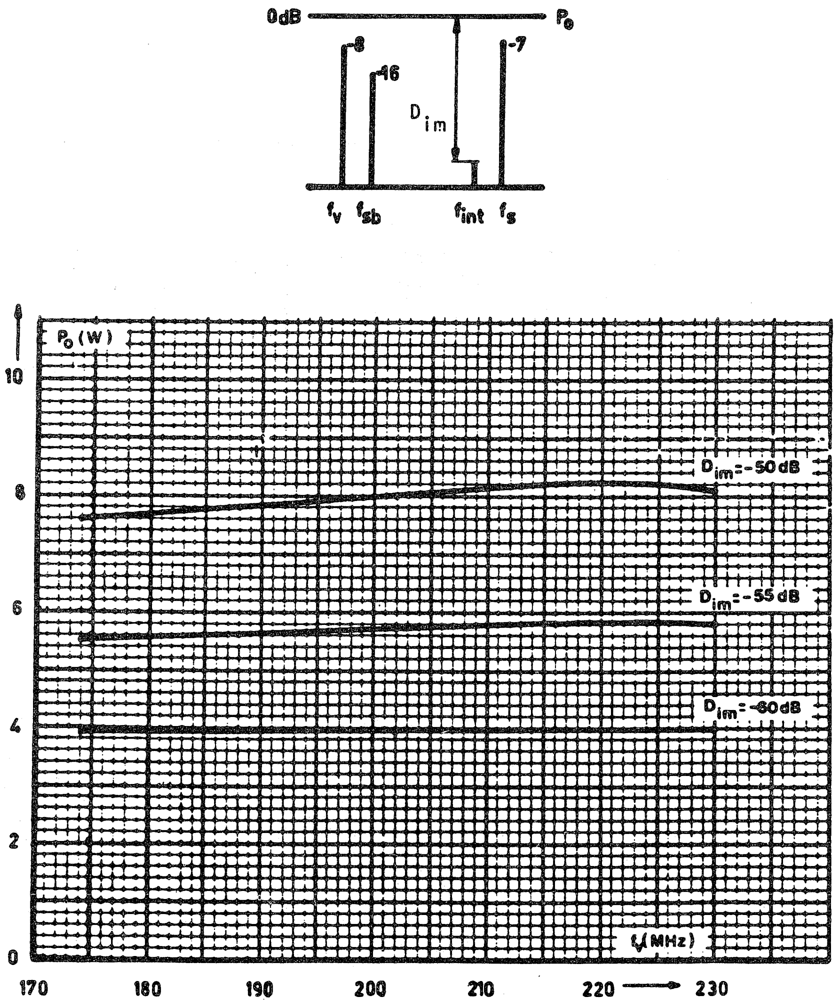
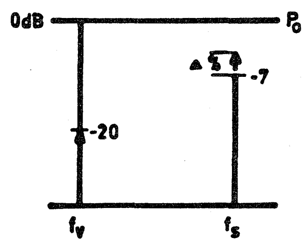


fig. 8: output power $P_{o_{sync}}$ of the BLV 30 amplifier



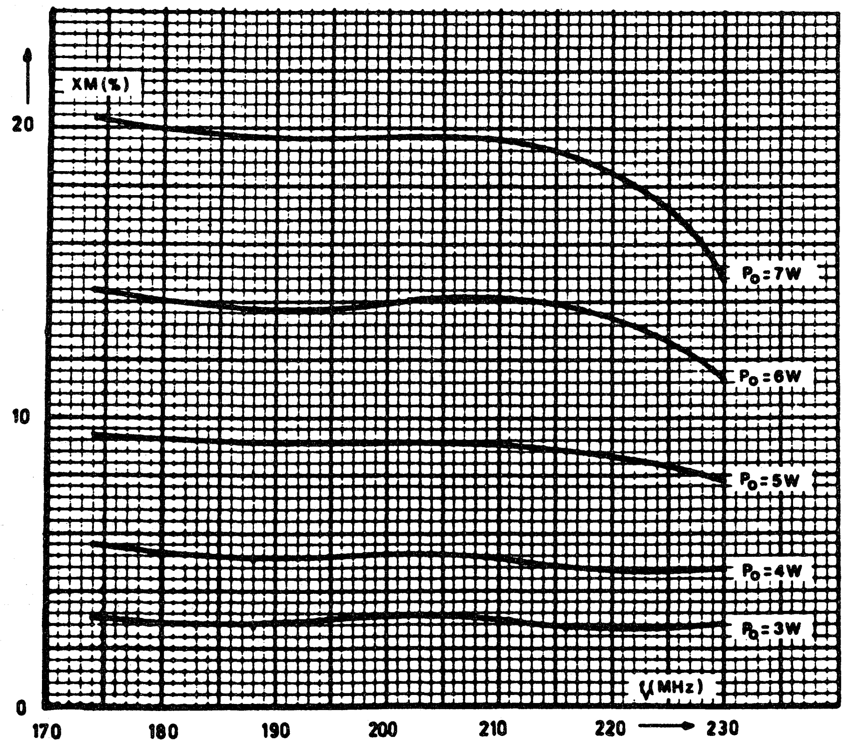


fig. 9: cross modulation of the BLV 30 amplifier

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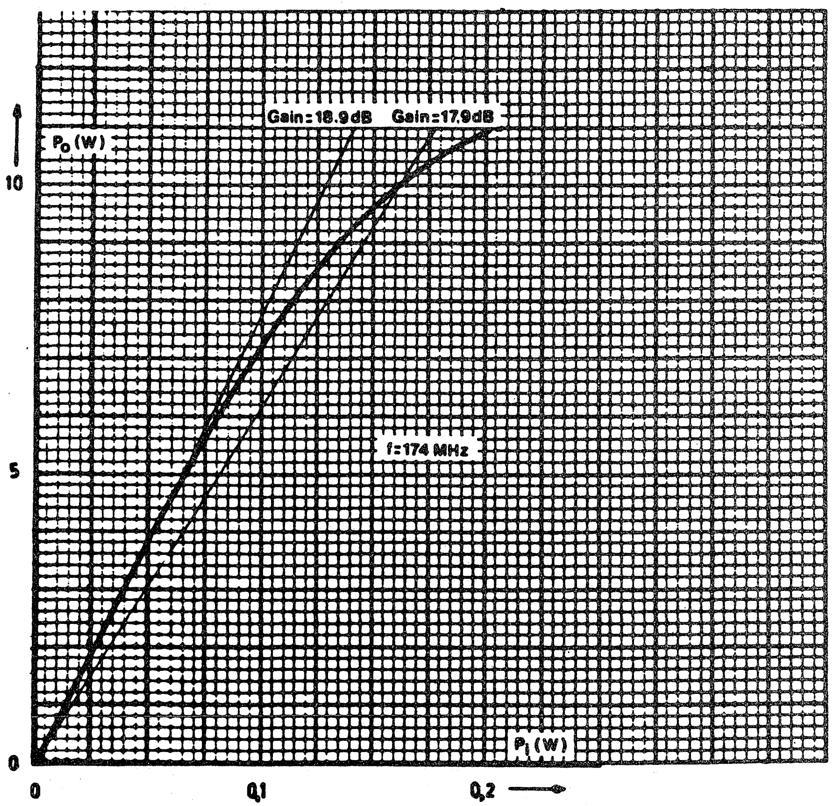

fig. 10: gain compression of the BLV 30 amplifier

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REPORT No: NCO 8103	AUTHOR: H. van Hees
PROJECT No:	DATE: 30.11.1981

TITLE

Wideband class AB power amplifier (174-230MHz) with two transistors BLV 36

ABSTRACT

For application in driver or final stages of TV transposers and transmitters in band III (174-230MHz) a wideband class AB power amplifier has been designed with two transistors BLV 36 coupled by means of 3dB-90° hybrids.

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Decision MAMO d.d: 29-12-81	AV	GV	EI	B		BL	
DATE: 29 DEC. 1981	MAMO:						

N.V. PHILIPS SEMICONDUCTORS APPLICATION LABORATORY
NIJMEGEN – THE NETHERLANDS

REPORT No: NCO 8103
PROJECT No:
AUTHOR: H. van Hees
DATE: 30.11.1981

TITLE

Wideband class AB power amplifier (174-230MHz) with two transistors BLV 36

SUMMARY

For application in TV transposers and transmitters a wideband power amplifier has been designed with two transistors BLV 36. It is a hybrid coupled amplifier for the frequency range 174-230MHz, with the BLV 36 operating in class AB. The main properties of the amplifier are gathered in the table below.

band 3: 174-230MHz	
DC setting	I_{CZ} = 2x 150mA V_{CE} = 28V
gain at P_{out} = 200 Watt	11.5dB ± 0.3dB
gain at 1dB gain compression	⩾ 10.5dB
P_{out} at 1dB gain compression	> 260 Watt
efficiency at P_{out} = 250 Watt	⩾ 45%

The applied p.c. board is double copper clad epoxy fibre glass ($\varepsilon_r \simeq 4.5$), thickness 0.8mm. The heatsink has forced air cooling. The supply voltage is 28V.

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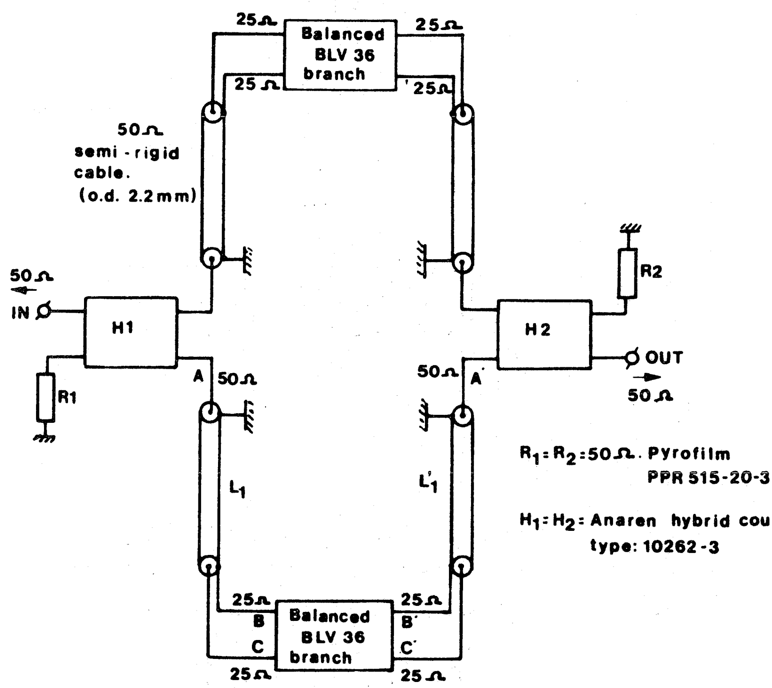
. Introduction

For application in TV transposers and transmitters for band 3 (174-230MHz) a wideband power amplifier has been designed with two transistors BLV 36. The BLV 36 is a balanced device, i.e. two identical transistor chips mounted in a single case, driven 180° out of phase. The case is an 8 lead envelope with a ceramic cap (NO 229). This transistor has been developed to operate with a class AB d.c. setting. The optimum quiescent current I_{CZ} = 150mA per chip and the V_{CE} = 28V.

2. Design of the amplifier

2.1 General remarks

The schematic line-up of a complete amplifier is given in fig. 1.



R1 = R2 = 50Ω. Pyrofilm PPR 515-20-3

H1 = H2 = Anaren hybrid coupler type: 10262-3

fig. 1: schematic line-up

The amplifier consists of two branches, both with a BLV 36 transistor, which are coupled by means of a wideband 3dB coupler at in- and output. Each BLV 36 has two input and two output circuits (one for each transistor chip). The baluns L_1 and L_1' connect the 25Ω balanced ports B and C to the unbalanced 50Ω port A. The phase shift between B and C amounts to 180^o. The balun is a 50Ω semi-rigid cable (2.2mm o.d.). The amplifier has been designed on epoxy glass fibre print material (ε_r = 4.5), thickness 0.8mm, copper clad on both sides.
To get a good contact between the upper and lower side, rivets have been used at several places and copper straps have been soldered alongside the edges of the board. Moreover, at the places where the emitters of the transistors have been grounded, contact has been established with the lower side of the print.

2.2 Properties of the BLV 36 class AB

The optimum d.c. setting of the BLV 36 for class AB operation is V_{CE} = 28V and a quiescent current I_{C_Z} = 150mA for each transistor chip. Typical gain, input impedance and optimum load impedance of a chip at P_o = 75W and V_{CE} = 28V are given in table 1.

table 1

freq. (MHz)	gain (dB)	input impedance (Ω)	load impedance (Ω)
174	12.23	0.75 + j0.92	2.59 + j0.85
181	12.01	0.83 + j0.99	2.51 + j0.82
188	11.81	0.92 + j1.06	2.43 + j0.78
195	11.63	1.05 + j1.13	2.35 + j0.74
202	11.47	1.21 + j1.19	2.27 + j0.69
209	11.33	1.43 + j1.21	2.19 + j0.63
216	11.21	1.70 + j1.19	2.12 + j0.85
223	11.10	2.04 + j1.04	2.05 + j0.51
230	11.02	2.38 + j0.70	1.97 + j0.44

The figures have been calculated with the aid of a large signal equivalent circuit for this device.

2.3 Output network

The 25Ω of the balun has to be transformed into the optimum load for half the transistor (one chip) which is given in table 1. This is done by means of an L-C output network which has been calculated according to ref. 1, and submitted to a computer optimization program. The BLV 36, being a balanced device, has two identical output circuits with a virtual ground inbetween. Fig. 2 shows this complete circuit.

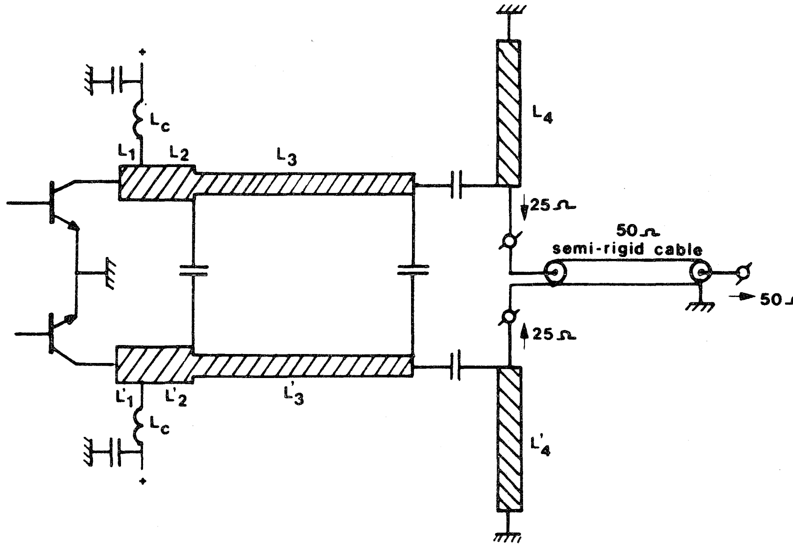


fig. 2: calculated output circuit

The collector biasing coil L_c has an active role in the transformation procedure. The coil outweights the imaginary part of the impedance at that point. The 50Ω semi-rigid coaxial cable (o.d. 2.2mm) is soldered atop the stripline L_4'.

2.4 Input network

The input impedance of one BLV 36 transistor chip (see table 1 page 3) has to be transformed into the 25Ω of the balun. At the same time the gain variation of 1.2dB has to be eliminated. This has been done by applying appropriate mismatch at the lower frequencies. This method is described in ref. 2. The L-C network also has been submitted to a computer optimization. Fig. 3 shows the input circuit for a BLV 36.

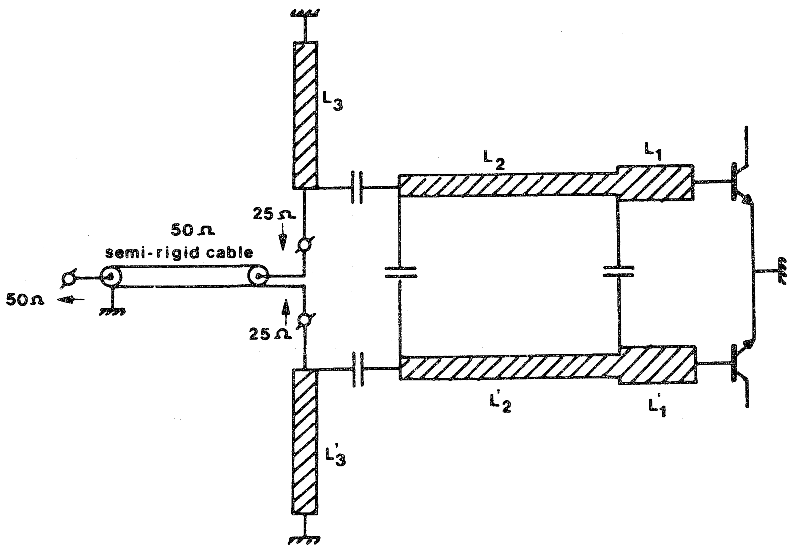


fig. 3: calculated input circuit

The 50Ω semi-rigid coaxial cable (o.d. 2.2mm) again has been soldered atop the stripline L_3'.

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3. Adjustment of the amplifier

3.1 Output circuit

In the previous section the theoretical design of an amplifier branch is given. To obtain maximum output power the BLV 36 should see the optimum load impedance, given in table 1 page 3. Practical tuning of the output circuit is established by replacing the BLV 36 by a dummy, consisting of a resistor and a capacitor in parallel, representing the complex conjugate of the optimum load impedance. The circuit is tuned for minimum reflection at the output terminal.
The applied dummy (soldered between the connection points of both collectors) has been calculated on 9Ω//82pF. Fig. 4 shows a typical curve return losses versus frequency.

3.2 Input circuit

After tuning the output circuit the dummy is replaced by a BLV 36 transistor. The two transistor chips get their d.c. setting (I_{CZ} = 150mA, V_{CE} = 28V) with the aid of a bias circuit* (each chip has its own bias circuit). Fig. 5 shows this circuit and the print layout. The gain of the amplifier branch is made sufficiently flat by tuning the input circuit. This is done under power sweep conditions. Fig. 6 shows the circuit of a tuned BLV 36 amplifier branch ; fig. 7 the gain versus frequency curve at an output power P_o = 100 Watt, plus the input return losses.

4. Hybrid coupled amplifier

In the previous section the tuning of one BLV 36 amplifier branch has been described. The gain has been made flat by means of mismatch at the input. By coupling two branches by means of 3dB-90^{o} hybrid couplers the influence of this mismatch at in- and output terminal is eliminated. These couplers reduce the return losses to at least 20dB. The reflected power is absorbed by the 50Ω termination resistor at the isolated port (see fig. 1 page 2).

* See appendix (page R 20).

Fig. 8 shows the p.c. board of the complete amplifier. The layout is in fig. 9.

5. Measured performance

5.1 Gain versus frequency

Fig. 10 shows the measured gain as a function of frequency at three levels of the output power.

5.2 Gain versus output power

The measured gain as a function of output power at two frequencies is in fig. 11. The collector efficiency is better than 45%. Table 2 shows the d.c. current per transistor chip at an amplifier output power of 250 Watt at several frequencies. These figures have been measured with a current meter with a clip-on d.c. current probe (HP 428B).

table 2

freq. (MHz)	P_{out} = 250 Watt			
	MP 16-27 no. 4		MP 16-21 no. 2	
	I_1 (A)	I_2 (A)	I_3 (A)	I_4 (A)
174	5.3	5	4.6	5.6
180	5.4	4.9	4.4	5.3
190	5.2	4.7	3.9	4.8
200	4.8	4.5	3.6	4.4
210	4.4	4.1	3.6	4.1
220	4	3.9	3.9	4.2
230	3.9	3.9	4.5	4.6

5.3 Output power versus input power

Fig. 12 shows output power versus input power at 174MHz. The 1dB compression point occurs at $P_{out} \simeq 270$ Watt. The output power at 1dB compression at several frequencies throughout the band is shown in fig. 13. Gain compression most rapidly occurs at 174MHz.

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6. Conclusion

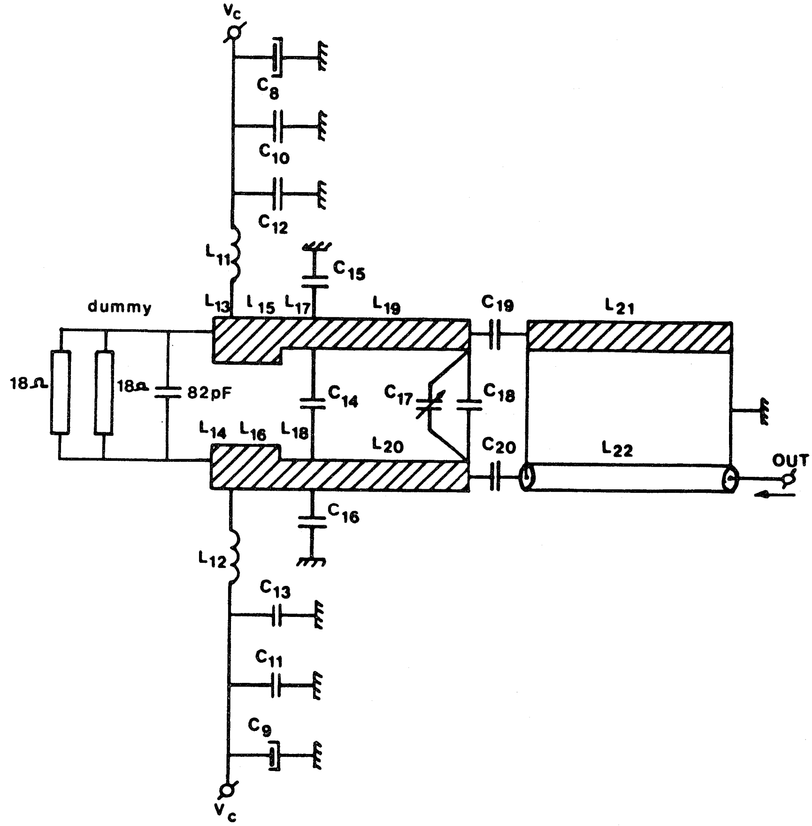
It is possible to build a class AB amplifier with two band 3 class AB transistors BLV 36. The main properties of this amplifier are:

band 3 174-230MHz	
DC setting BLV 36	I_{C_Z} = 2x 150mA V_{CE} = 28V
gain at P_{out} = 200 Watt	11.5dB $\pm$ 0.3dB
gain at 1dB gain compression	$\geqslant$ 10.5dB
P_{out} at 1dB gain compression	> 260 Watt
efficiency at P_{out} = 250 Watt	$\geqslant$ 45%

7. References

Ref. 1: G.L. Matthaei
Tables of Chebychev impedance transforming networks of low pass filter form.
Proc. of the IEEE, August 1964.

Ref. 2: O. Pitzalis Jr. and R.A. Gibson
Tables of impedance matching networks which approximate prescribed attenuation versus frequency slopes.
IEEE transactions on microwave theory and techniques, vol. MTT 119, no. 4, April 1971.



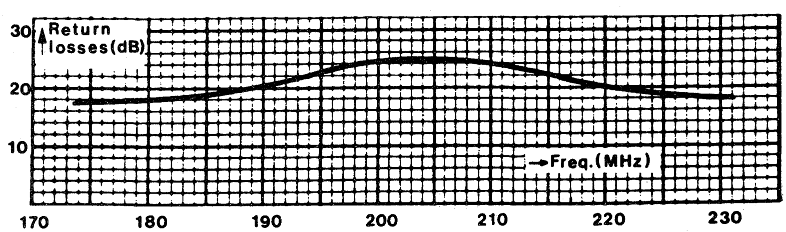
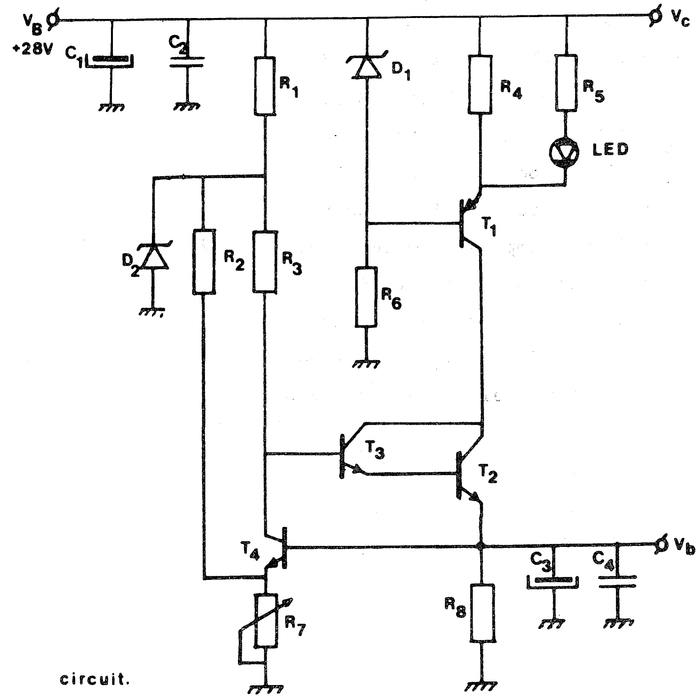
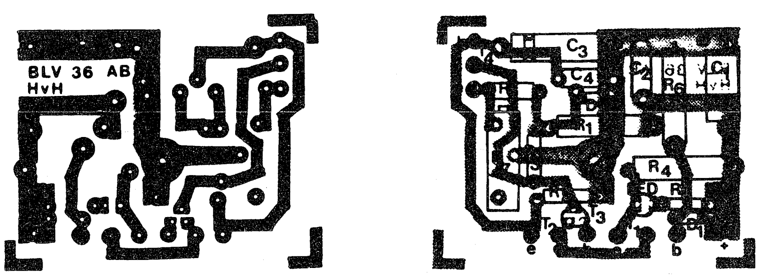


Fig. 4. Tuning the output circuit.



circuit.



print and lay-out. Scale 1:1

Fig.5. Bias-circuit.

Parts list bias circuit

R_1 = 1k2 carbon resistor CR 68

R_2 = 4k7 carbon resistor CR 25

R_3 = 390 carbon resistor CR 25

R_4 = 22Ω carbon resistor CR 68

R_5 = 39Ω carbon resistor CR 16

R_6 = 2k2Ω carbon resistor CR 68

R_7 = 500Ω trimpotentiometer, cat.no. 2122 350 00045

R_8 = 68Ω carbon resistor CR 37

C_1 = 47µF/40V elco, cat.no. 2222 030 37479

C_2 = C_4 = 100nF/250V, metallised film, cat.no. 2222 352 45104

C_3 = 100µF/40V elco, cat.no. 2222 031 37101

T_1 = BD 234

T_2 = BD 233

T_3 = BC 546

T_4 = BD 139

D_1 = BZX 79 C/6V8

D_2 = 1N 825

LED = CQY 94

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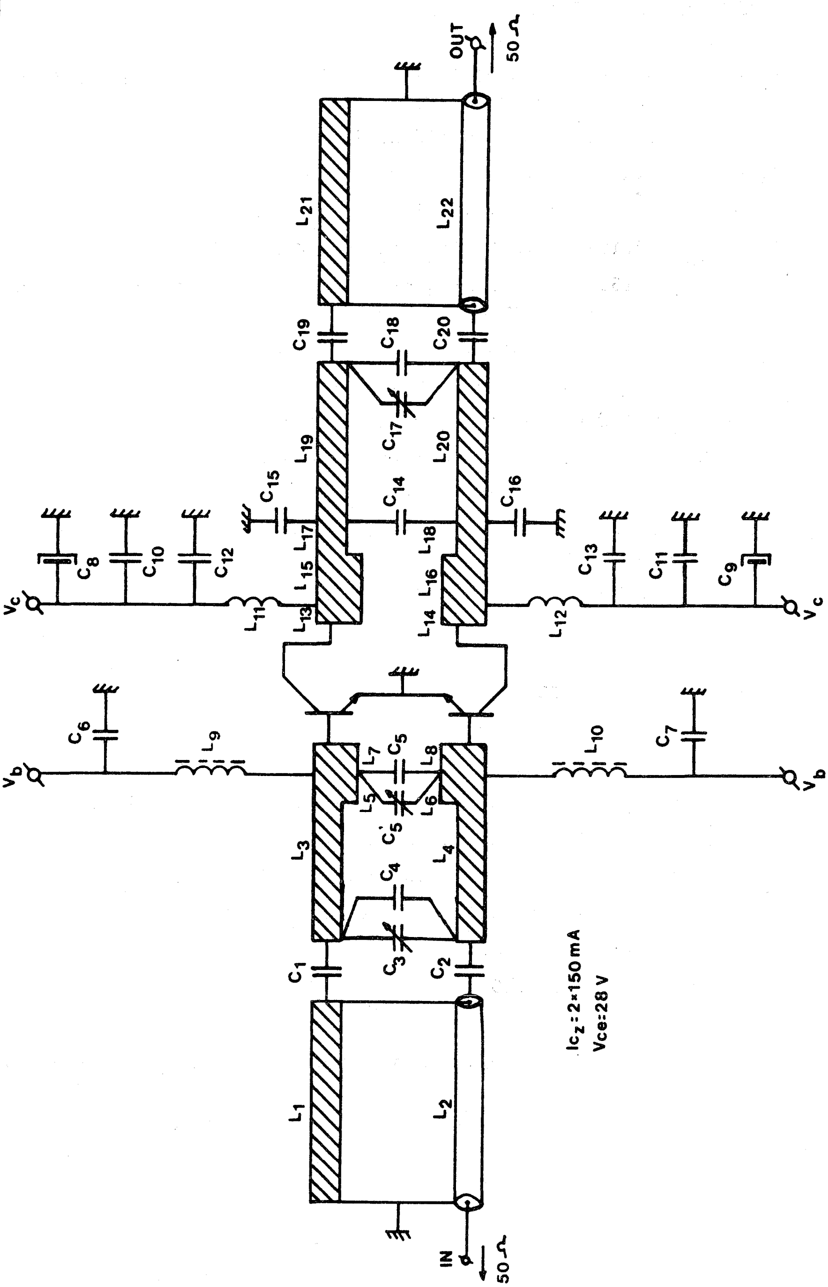

Fig.6. Circuit of a BLV 36 branch.

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Parts list of a BLV 36 branche

L_1 = 41Ω stripline, w = 2mm, l = 101mm

L_2 = 50Ω semi-rigid coaxial cable, d = 2.2mm, l = 101mm soldered on a 41Ω stripline, w = 2mm, l = 101mm

$L_3 = L_4$ = 41Ω stripline, w = 2mm, l = 30mm

$L_5 = L_6$ = 31Ω stripline, w = 3mm, l = 5mm

$L_7 = L_8$ = 31Ω stripline, w = 3mm, l = 4.4mm

$L_9 = L_{10}$ = 0.1μH choke Philips cat.no. 4322 057 01071

$L_{11} = L_{12}$ = 15.1nH; 1 winding of enamelled Cu wire, Ø 1.5mm, int. diam. 8.3mm, leads 5mm

$L_{13} = L_{14}$ = 31Ω stripline, w = 3mm, l = 3mm

$L_{15} = L_{16}$ = 31Ω stripline, w = 3mm, l = 12.5mm

$L_{17} = L_{18}$ = 41Ω stripline, w = 2mm, l = 3mm

$L_{19} = L_{20}$ = 41Ω stripline, w = 2mm, l = 34mm

L_{21} = 41Ω stripline, w = 2mm, l = 113mm

L_{22} = 50Ω semi-rigid coaxial cable, d = 2.2mm, l = 113mm, soldered on a 41Ω stripline, w = 2mm, l = 113mm

$C_1 = C_2$ = 30pF chip, ATC 100 B

$C_3 = C_{17}$ = 2-18pF film dielectric trimmer, Philips cat.no. 2222 809 05003

C_4 = 6.2pF chip, ATC 100 B

C_5 = 91pF chip, ATC 100 B

C_5' = 1.5-9pF film dielectric trimmer, Philips cat.no. 2222 809 05002

$C_6 = C_7 = C_{12} = C_{13}$ = 680pF chip, Philips cat.no. 2222 852 13681

$C_8 = C_9$ = 47μF elco, Philips cat.no. 2222 030 37479

$C_{10} = C_{11}$ = 100nF chip, Philips cat.no. 2222 854 48104

$C_{14} = C_{19} = C_{20}$ = 56pF chip, ATC 100 B

$C_{15} = C_{16}$ = 82pF chip, ATC 100 B

C_{18} = 15pF chip, ATC 100 B

P.C. board material is epoxy fibre-glass ($\varepsilon_r \simeq 4.5$), thickness 0.8mm

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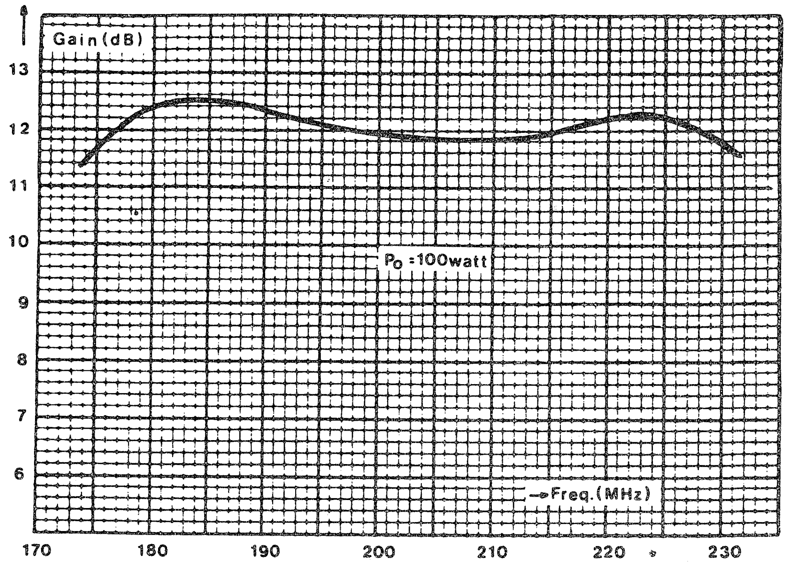

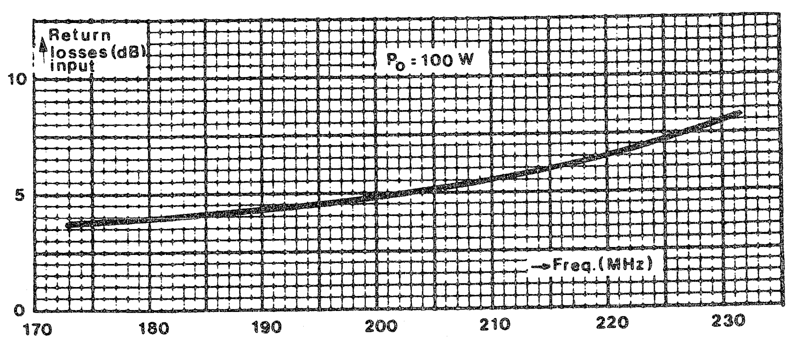

Fig.7. Gain and input return losses of a BLV 36 branch.

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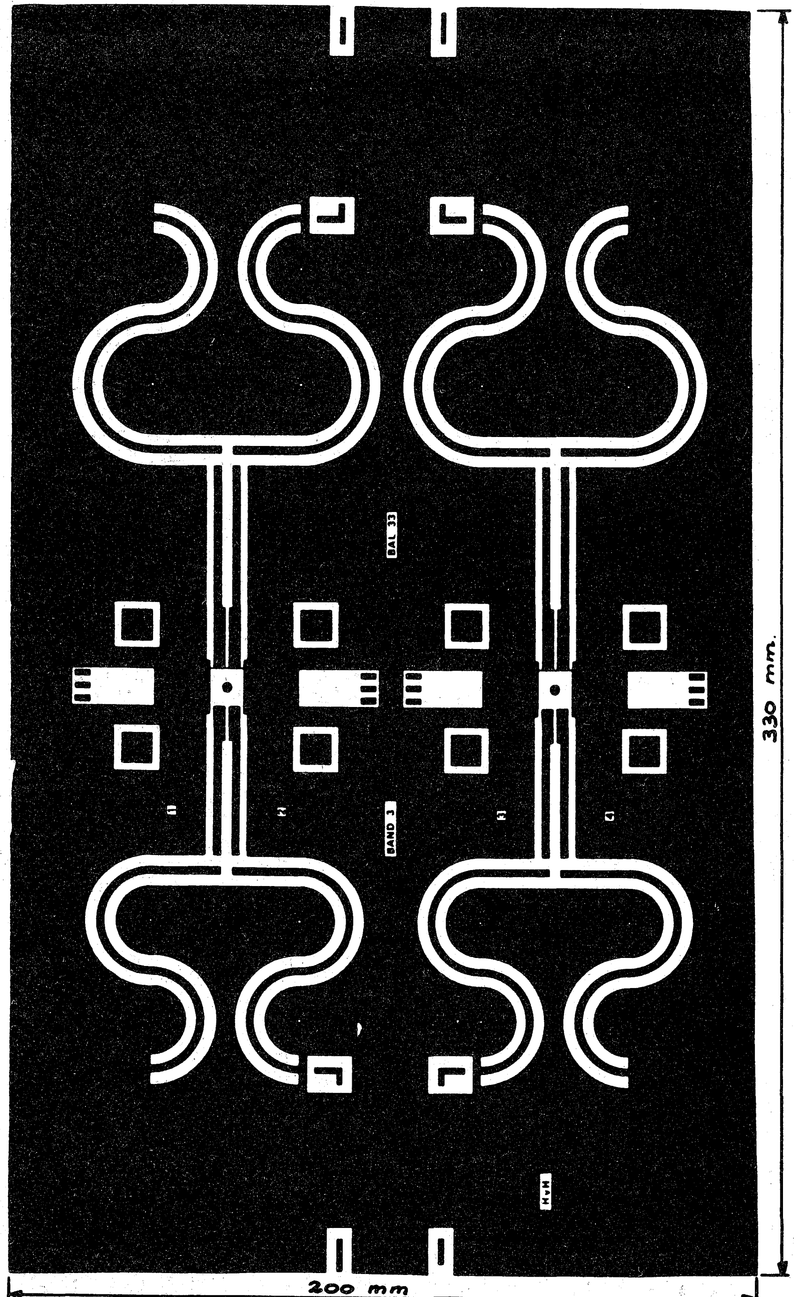

Fig.8. Amplifier p.c. board.

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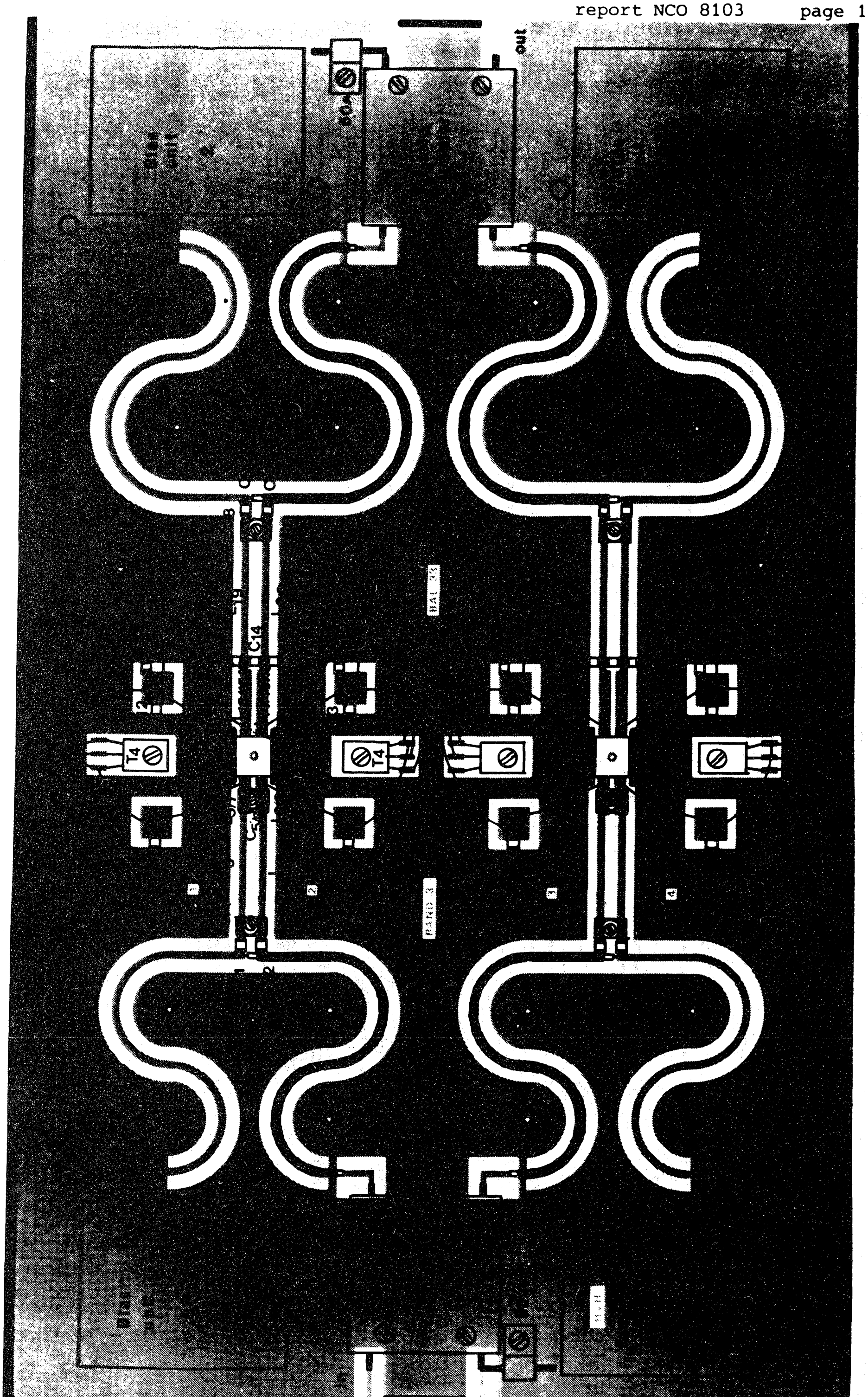


Fig.9. Amplifier lay-out.

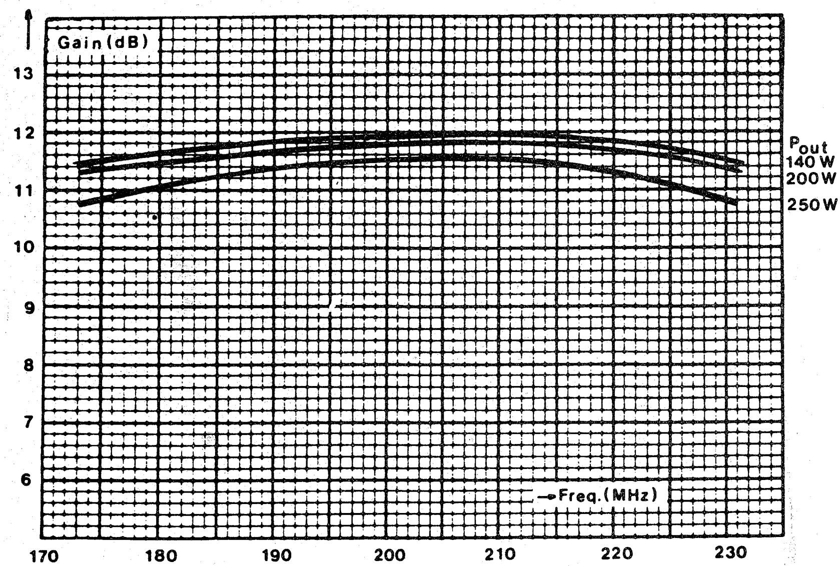


Fig. 10. Gain versus frequency.

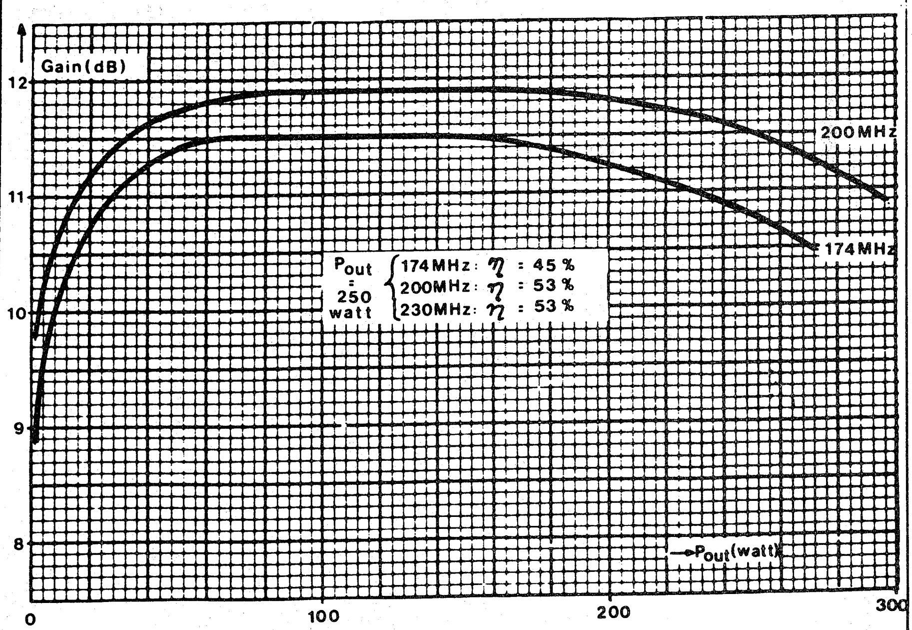


Fig. 11. Gain versus output power.

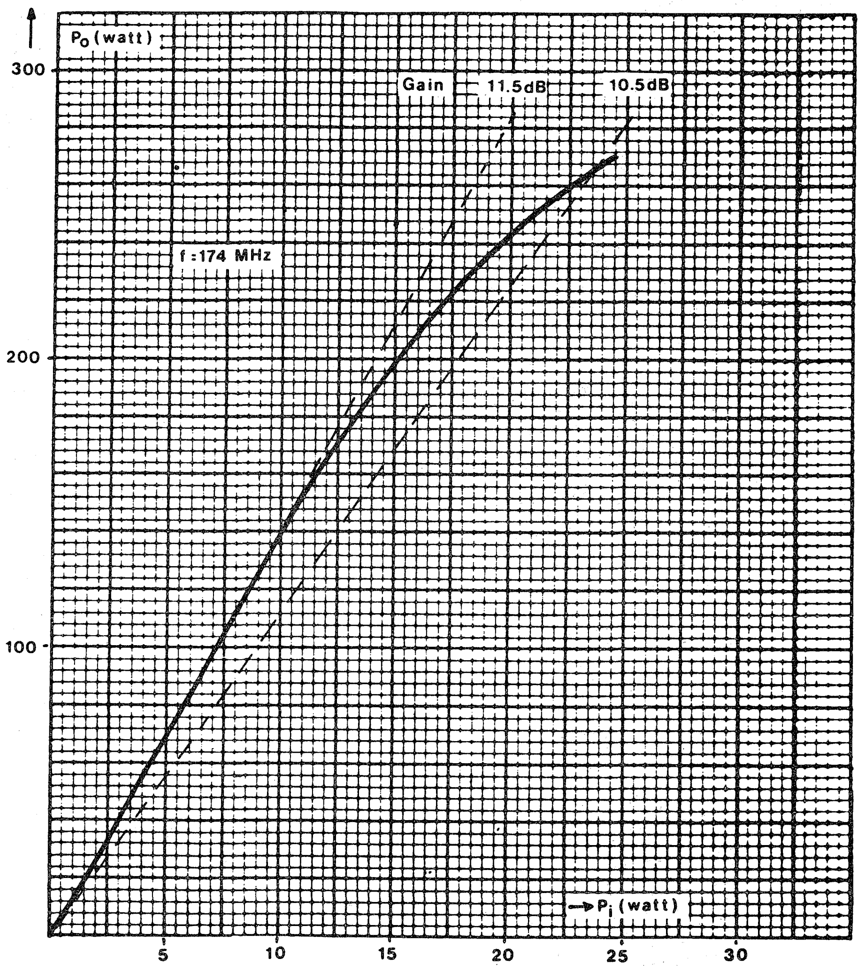


Fig.12. Pout versus Pin.

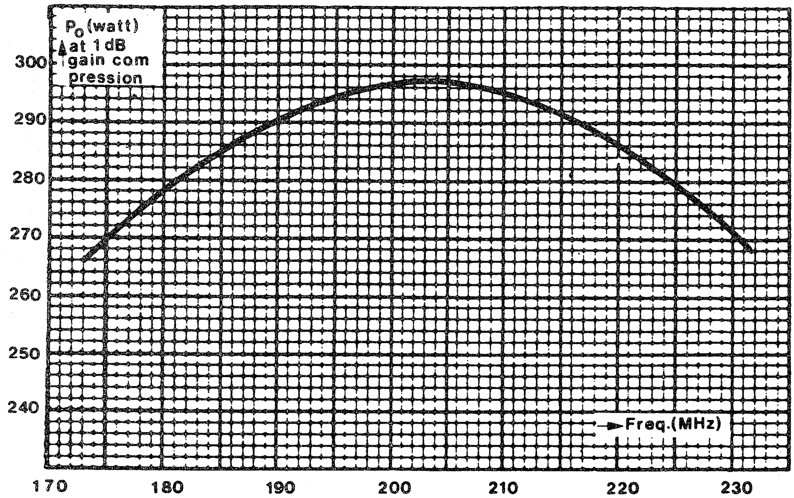
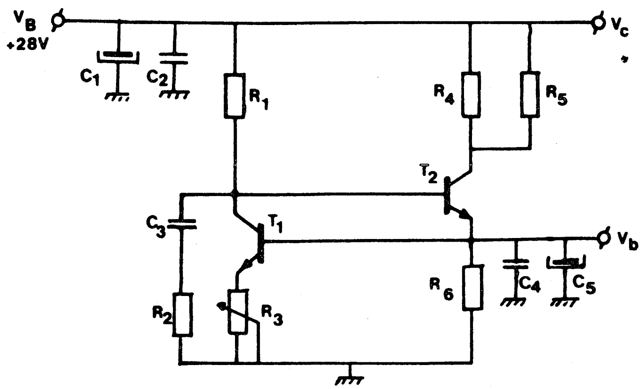


Fig. 13. Pout at 1dB gain compression.

Appendix

The applied bias circuit is an universal class AB bias circuit, commonly used in our development laboratory for testing a wide range of different transistor types. Therefore it has some features, e.g. protection against short circuit in the h.f. transistor, the possibility of varying V_b between 0.4 Volt and 1 Volt and the collector voltage V_C between 12 Volt and 55 Volt.
A more simple, but adequate bias circuit for the 2x BLV 36 amplifier is shown in the figure below. Transistor T_1 should follow the temperature variation of the BLV 36. The R_2-C_3 combination prevents the bias circuit from oscillating.



Parts list

R_1 = 820Ω carbon resistor CR 68

R_2 = 10Ω carbon resistor CR 25

R_3 = 10Ω cermet potentiometer cat.no. 2122 350 00056

R_4 = R_5 = 150Ω enamelled wire wound cat.no. 2322 330 32151

R_6 = 22Ω carbon resistor CR 25

C_1 = 47µF/40V elco Philips cat.no. 2222 030 37479

C_2 = C_3 = C_4 = 100nF/250V metallised film, cat.no. 2222 352 45104

C_5 = 100µF/40V elco Philips cat.no. 2222 031 37101

T_1 = BD 135

T_2 = BD 233

A I

N.V. PHILIPS SEMICONDUCTORS APPLICATION LABORATORY
NIJMEGEN – THE NETHERLANDS

REPORT No: NCO 8201	AUTHOR: G.Lukkassen
PROJECT No:	DATE: 1982-01-07

TITLE

Construction of the BLV 57 wideband amplifier (470-860 MHz).

ABSTRACT

Additional to report NCO 8101 more information is given about the construction of a BLV 57 band 4/5 linear power amplifier for T.V. transposer applications.
Attention has been payed to good mechanical and electrical contacts, and a low thermal resistance by means of a forced air-cooled heatsink.

appr. J.Tuil

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1. Introduction

In the application report NCO 8101 two amplifiers for band 4/5 with BLV 57 transistors have been described.
Reactions on this report proved the necessity to give more information about the construction of these amplifiers.
This construction has been based on a heatsink with a PC-board at the upper side and the bias circuits and a forced air-cooling at the lower side.

2. Printed circuit board

In the PC-board rectangular holes have been made to mount the BLV 57 transistors on the heatsink. For fastening of the PC-board on the heatsink by means of screws, 7 holes of 3.1 mmØ and for the fastening of the hybrid couplers 8 holes of 2.6 mm Ø have been made on the indicated places (see fig. 1).
Hereby has been taken into account the use of Anaren hybrid couplers, type 10264-3, suited for the frequency range of 500-1000 MHz.
Because the 2 bias units have been situated at the lower side of the heatsink, the connections from these units to the circuit take place through the PC-board and the heatsink.
For this purpose 9 holes of 2 mm Ø are necessary (4 collectors, 4 bases and 1 ground).
To make a good ground contact between the upper and the lower side of the PC-board the following measures have been taken:

- On 8 spots rivets have been used and soldered at both sides to the metallisation of the PC-board. The holes of 2 mm Ø, needed for these rivets, have been situated as indicated in fig. 1.
- Copper straps with a thickness of 0.2 mm have been soldered at all edges of the PC-board.

- A good emitter to ground contact has been achieved by soldering 8 copper straps from the upper to the lower side of the PC-board on the spots of each emitter lead.
- The input connector and the output connector have been screwed to the heatsink but the ground also has been soldered to the PC-board.

3. Heatsink

For the BLV 57 amplifiers, described in report NCO 8101, a blackened heatsink of Seifert Electronic, type KL-117 with a length of 191 mm has been used (see fig. 2).
At the lower side forced air-cooling has been applied with a fan trade mark Etri, type 99XU 01-81 with an air displacement of 16 litres per second (see fig. 3).
By applying this air-cooling the thermal resistance decreased from 0.5 $^{o}C/_{W}$ to 0.2 $^{o}C/_{W}$.

4. Mechanical machining of the heatsink

The raised edges at the top side of the heatsink have been removed because the printed circuit board has a width of 113 mm (see fig. 2).
To fit the heatsink to the PC-board the following machinings have been carried out:

- Rectangular holes of 2.8 mm deep have been mould in the heatsink because the transistor leads have to be soldered on the PC-board.
  Also it was necesarry to make savings of 4 mm wide and 0.6 mm deep at the positions of the straps on the PC-board. The transistors have been fastener with M 2.5 screws in the heatsink (see fig. 4).
- To achieve that the PC-board lays tight to the heatsink also savings have been made in the heatsink on the spots of the 8 rivets through the PC-board.

- For fastening the PC-board on the heatsink on 7 places holes with M3 screwthread have been made in the top side of the heatsink, corresponding with the indicated holes in the PC-board.
- The two hybrid couplers also have been fastened in the heat-sink with screws through the PC-board. Therefore 8 holes have been made with M 2.5 screwthread, corresponding with the PC-board holes.
- The input and output connectors have been fastened to the heatsink with M3 screws. The mid contact of each connector makes contact with the PC-board.

5. Conclusions

With the construction of the BLV 57 amplifiers a good thermal resistance (0.2 $^{o}C/_{W}$) has been achieved by means of a forced air-cooling.

Attention has been payed to a good mechanical contact between heatsink and printed circuit board and a good ground contact on the PC-board by means of rivets and straps at the edges and under the emitter leads.

PHILIPS

Electronic components and materials

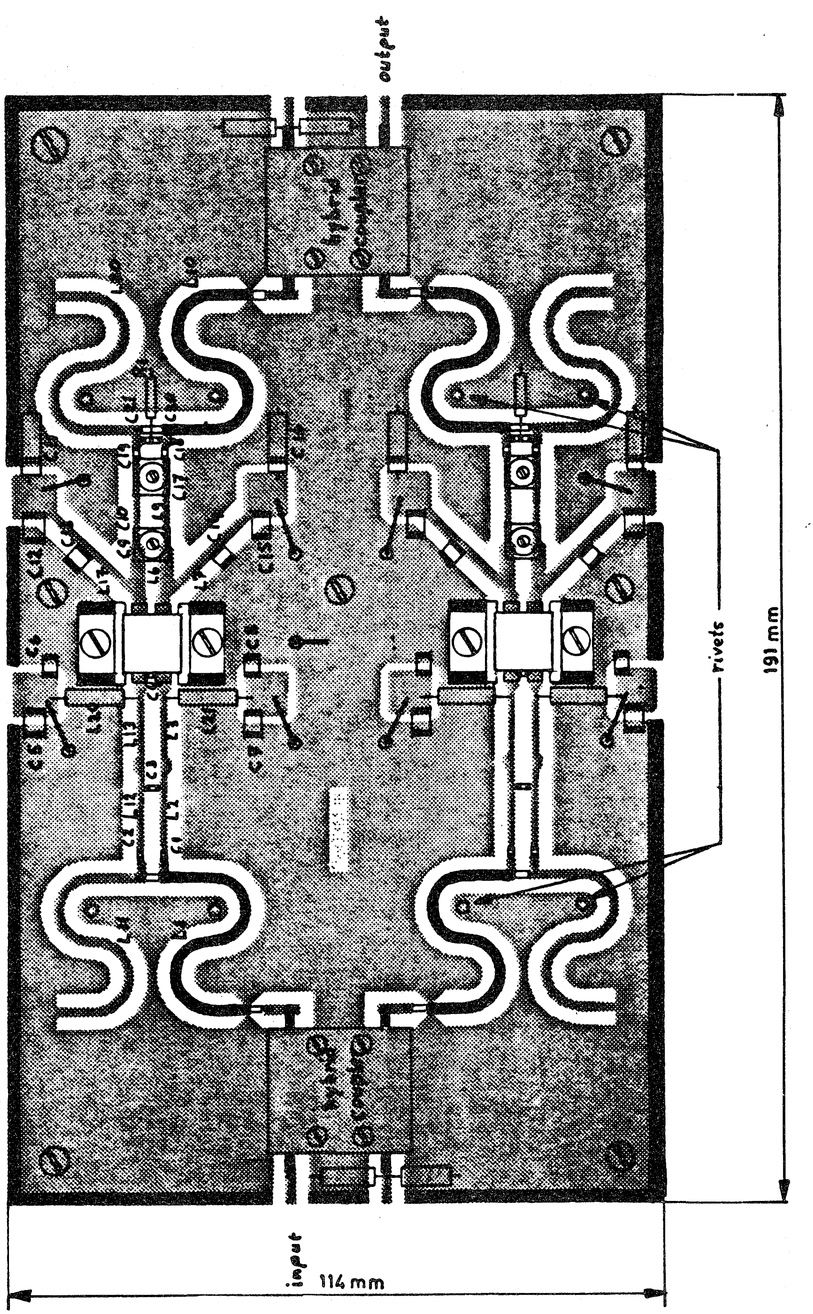

Fig. 1 Printed circuit board

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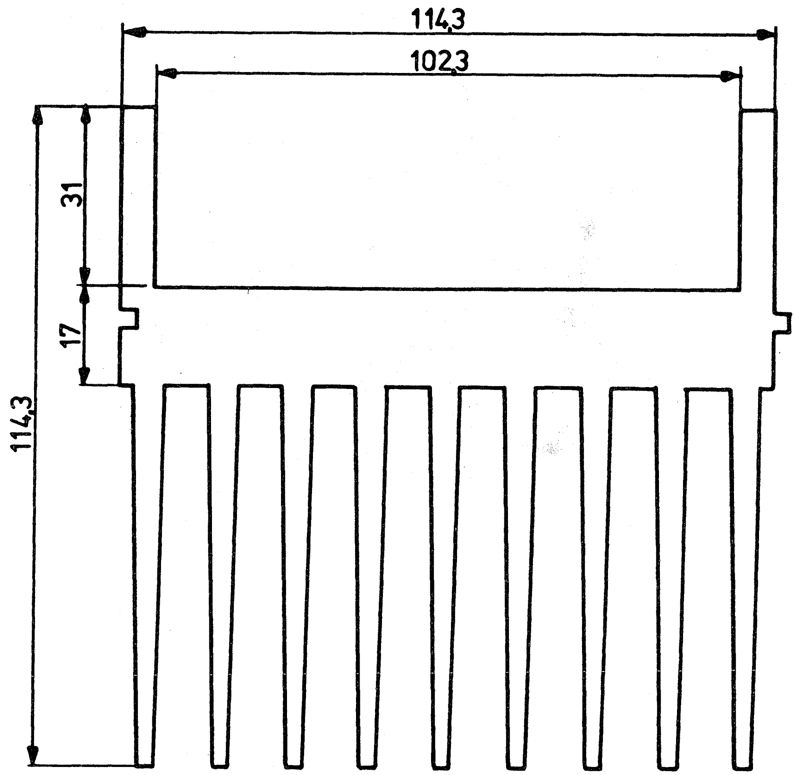

Serie KL-117

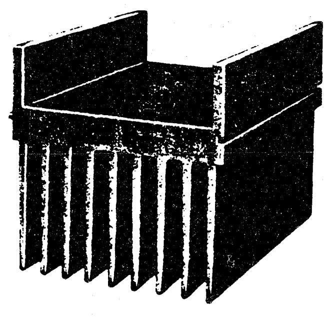

Fig. 2 Heatsink

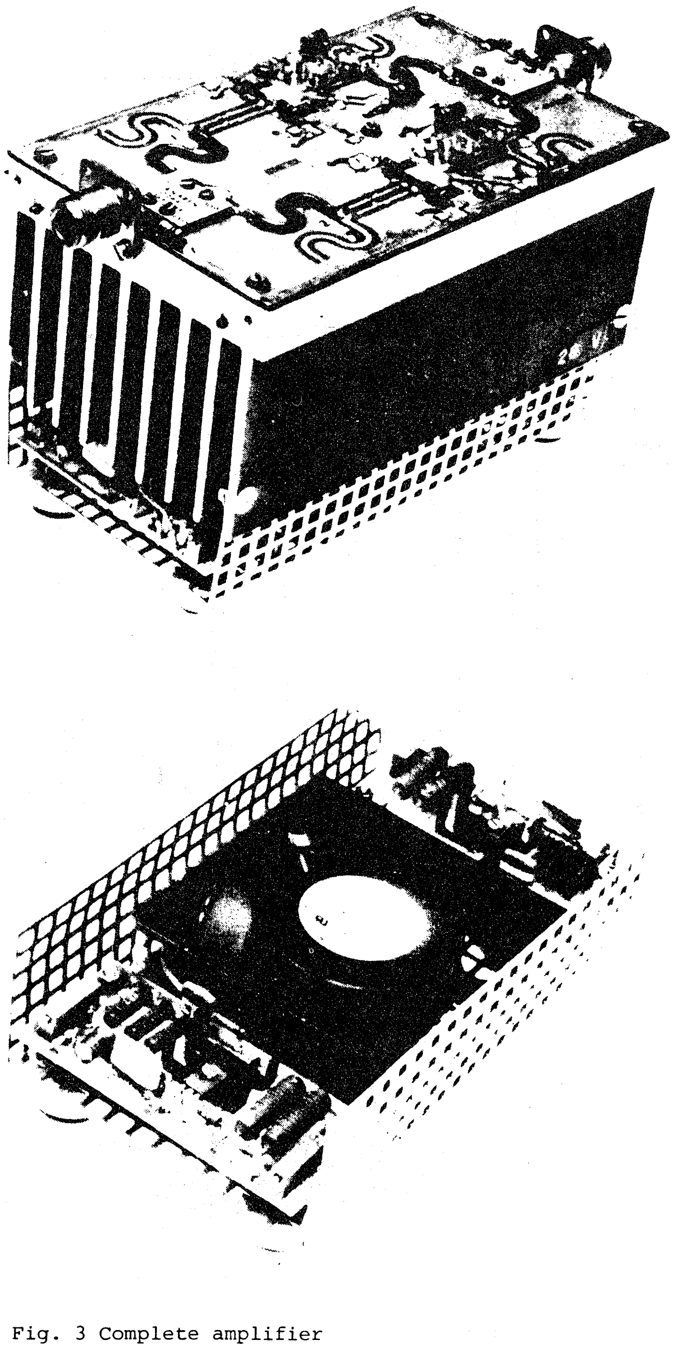

Fig. 3 Complete amplifier

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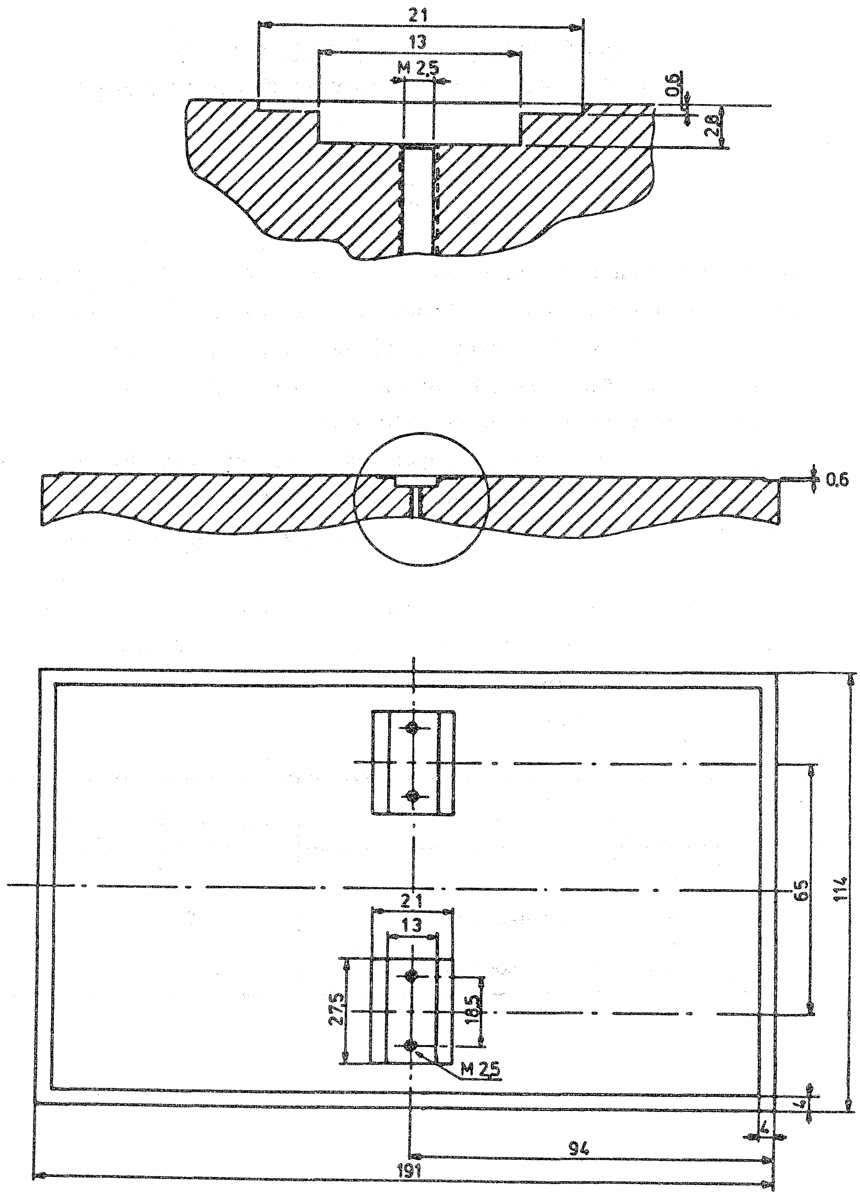

Fig. 4 Heatsink savings

Semiconductor application laboratory SAL Nijmegen - The Netherlands

RNR-1-092-1982-AS/ NCO 8205

By: G.Lukkassen

Applikatie Laboratorium

Nijmegen, 1982-04-19

A WIDEBAND CLASS AB HYBRID COUPLED AMPLIFIER (470-860 MHz) WITH TWO BALANCED TRANSISTORS BLV 57

SUMMARY

For application in TV transmitters in band 4/5 a wideband linear power amplifier has been designed with two balanced transistors BLV 57 in a class AB DC-setting (V_{CE}=25V and I_{CZ}=2x100mA).
A class A amplifier designed around the BLV 57 has been described in reports NCO 8101 and NCO 8201.
The results of the class AB input and output circuit calculations are about similar to the results of the class A application.
Therefore the p.c.-board design of the class A-amplifier can be used.
The applied circuit board is a double copper clad PTFE fibreglass print with an ε_r=2.74 and a thickness of 1/32 inch.
The heatsink has a forced air cooling.
The main results are:

DC-setting	I_{CZ}=4x100mA, V_{CE}=25V
Gain at P_{out}=5Watt	≥ 6 dB
P_{out} at 1dB gain compression	≥ 42.5W
Efficiency at 1dB gain compression	≥ 45 %

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1. INTRODUCTION

The BLV 57 is a balanced transistor in an 8 lead envelope (SOT 161) for class A operation in TV-transposers for band 4/5.
A class A amplifier, designed around two transistors BLV 57, has been described in report NCO 8101 and the construction of this amplifier in report NCO 8201.
Because there is also a typical class AB specification a wide-band power amplifier has been designed around two transistors BLV 57 in class AB.
The quiescent current I_{CZ}=100mA per chip and the V_{CE}=25V.

2. DESIGN OF THE AMPLIFIER

2.1. General remarks

The schematic line-up of the complete amplifier is given in fig.1.

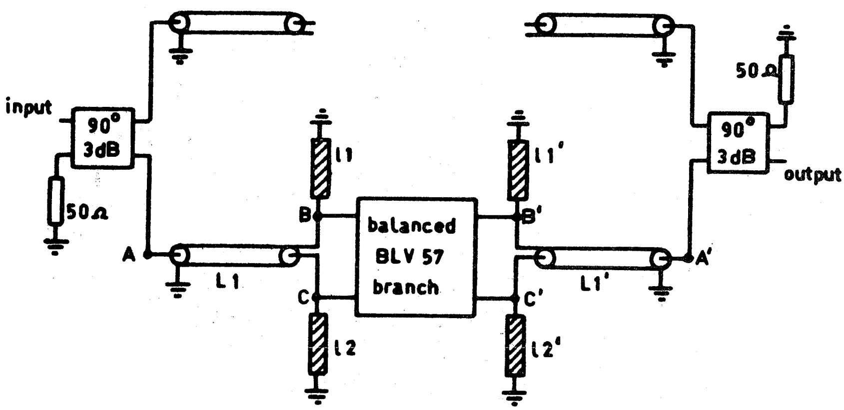


Fig.1: Schematic line-up

The amplifier consists of two branches, both with a BLV 57 transistor, which are coupled by means of a wide-band 3dB-90° coaxial hybrid coupler at the input and output.
Each BLV 57 has 2 input circuits and 2 output circuits (one for each chip) connected to a coax balon (L_1 and L_1') which connects the 25Ω balanced ports B and C to the unballanced 50Ω port A.

The phase-shift between B and C is 180^{o}.
The p.c.-board design, the material and the construction of the amplifier is equal to the class A amplifier described in the reports NCO 8101 and NCO 8201.

2.2. Bias circuit

Each transistor has its own bias unit to obtain a stable DC-setting for class AB operation (see fig.2). This bias unit anables a stable adjustment of the collector currents of the BLV 57 by means of potentiometer R_2.
To follow the temperature variation of the BLV 57 the transistor T_1 has been situated on the heatsink near to the HF-transistor for a good thermal contact (see fig.5).

2.3. Some properties of the BLV 57

The optimum DC-setting of the BLV 57 for class AB operation is V_{CE}=25V and a quiescent current of I_{CZ}=100mA for each transistor chip. The typical gain, input and load impedance of a half BLV 57 (one chip) are given in table I.
These figures have been calculated with the aid of a large signal equivalent circuit (P_o=17.5W).

Frequency (MHz)	Gain (dB)	Input impedance (Ω)	Output impedance (Ω)
400	11.95	1.21 + j 1.71	10.52 + j 4.04
500	10.29	1.24 + j 2.53	9.06 + j 4.02
600	9.01	1.28 + j 3.32	7.70 + j 3.63
700	7.99	1.36 + j 4.11	6.50 + j 2.98
800	7.17	1.49 + j 4.93	5.49 + j 2.13
900	6.48	1.68 + j 5.81	4.67 + j 1.17

Table I

../6

2.4. Input and output circuit

The calculation of the input and output circuit is the same as described in NCO 8101 chapter 2.4. and 2.5.
The results are about similar to the results of the class A application, making it possible to apply the same p.c.-board design.
The tuning of the output circuit is also as described in report NCO 8101. The dummy now consists of a 30Ω resistance in parellel with an 8.2pF capacitance.
In fig.3 the return losses at the output of one branch are given after tuning the output circuit with the help of this dummy.To achieve a sufficiently flat gain the capacitance of C_3 and C_4 and also the position of C_3 (see fig.4) can be optimized in a sweep set-up with a constant output power of 5W. The position of C_4 and C_5 is close to the ceramic cap of the BLV 57.

3. HYBRID COUPLED AMPLIFIER

As mentioned in chapter 4 of NCO 8101 the two branches are coupled by means of 3dB-90^o hybrid couplers. Fig.5 gives the p.c.-board of the complete amplifier and Fig.6 the lay-out.

4. MEASURED PERFORMANCE

4.1. Gain and return losses

Fig.7 shows the gain and input return losses as a function of the frequency at a constant output power P_o=5W. The gain varies from 6 to 6.9dB. The input return losses are at least 12dB.
Fig.8 shows the gain versus output power at the frequencies 500MHz and 800 MHz. The increase of gain at low power level can be reduced at the cost of the average gain level by decreasing the quiescent current.

../7

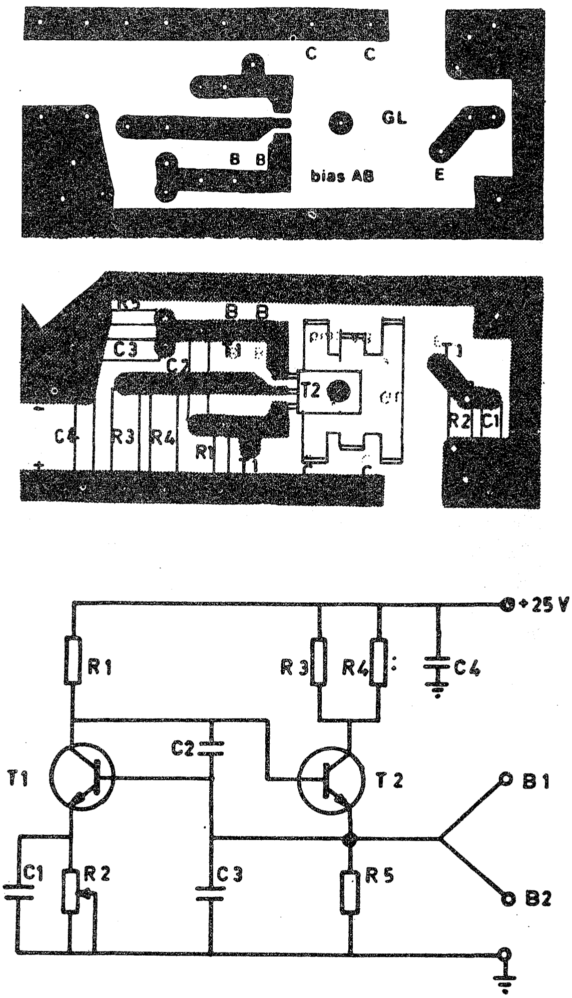
4.2. Output power

Fig.9 shows the output power as a function of the frequency at 1dB gain compression. The output power is at least 42.5Watt and above 530 MHz between 50 and 60 Watt.
The average efficiency at 1dB gain compression is 50%.

5. CONCLUSION

This report shows that it is possible to operate the class A transistor BLV 57 with a class AB DC-setting in a hybrid coupled wideband amplifier (470-860MHz) with good performances.
The main properties of the amplifier are:

BLV 57 band 4/5 amplifier	
DC-setting	I_{CZ}=4x100mA,V_{CE}=25V
Gain at P_{out}=5W	$\geqslant$ 6dB
P_{out} at 1dB gain compression	$\geqslant$ 42.5 Watt
Efficiency at 1dB gain compression	$\geqslant$ 45%



$C_1=C_2=C_3=C_4=150nF$, metallised film capacitor,cat.no.2222 352 45154

$R_1=1500\ \Omega$, CR 25 type, cat.no. 2322 211 13152

$R_2=10\Omega$, cermet potentiometer, cat.no.2122 350 00056

$R_3=R_4=120\Omega$, CR 52 type, cat.no. 2322 213 13121

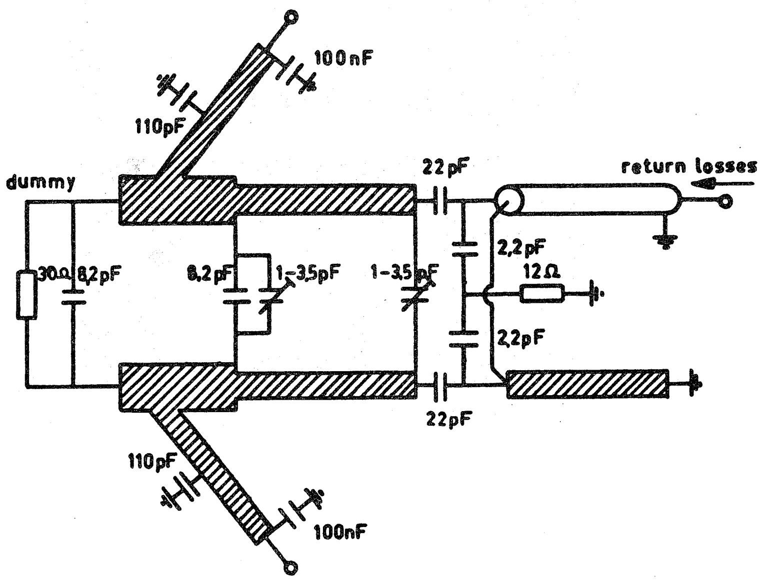
$R_5=150\Omega$, CR 25 type, cat.no. 2322 211 13151

$T_1=T_2=$ BD 139

Fig.2 Bias circuit and lay-out

PHILIPS

Electronic components and materials



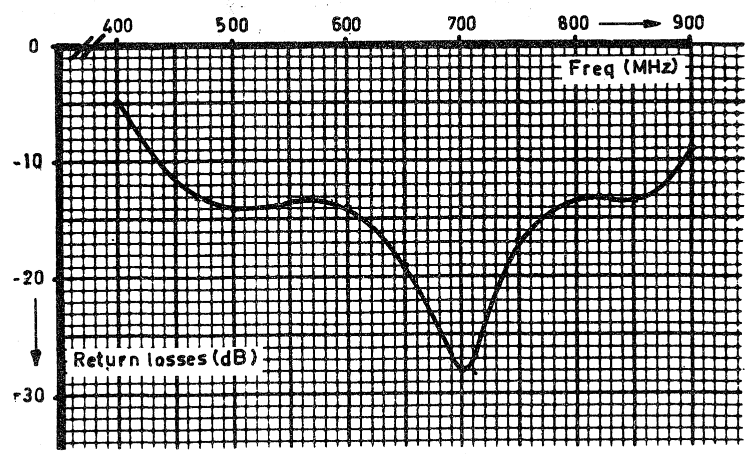


Fig.3 Tuning of the output circuit

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Electronic components and materials

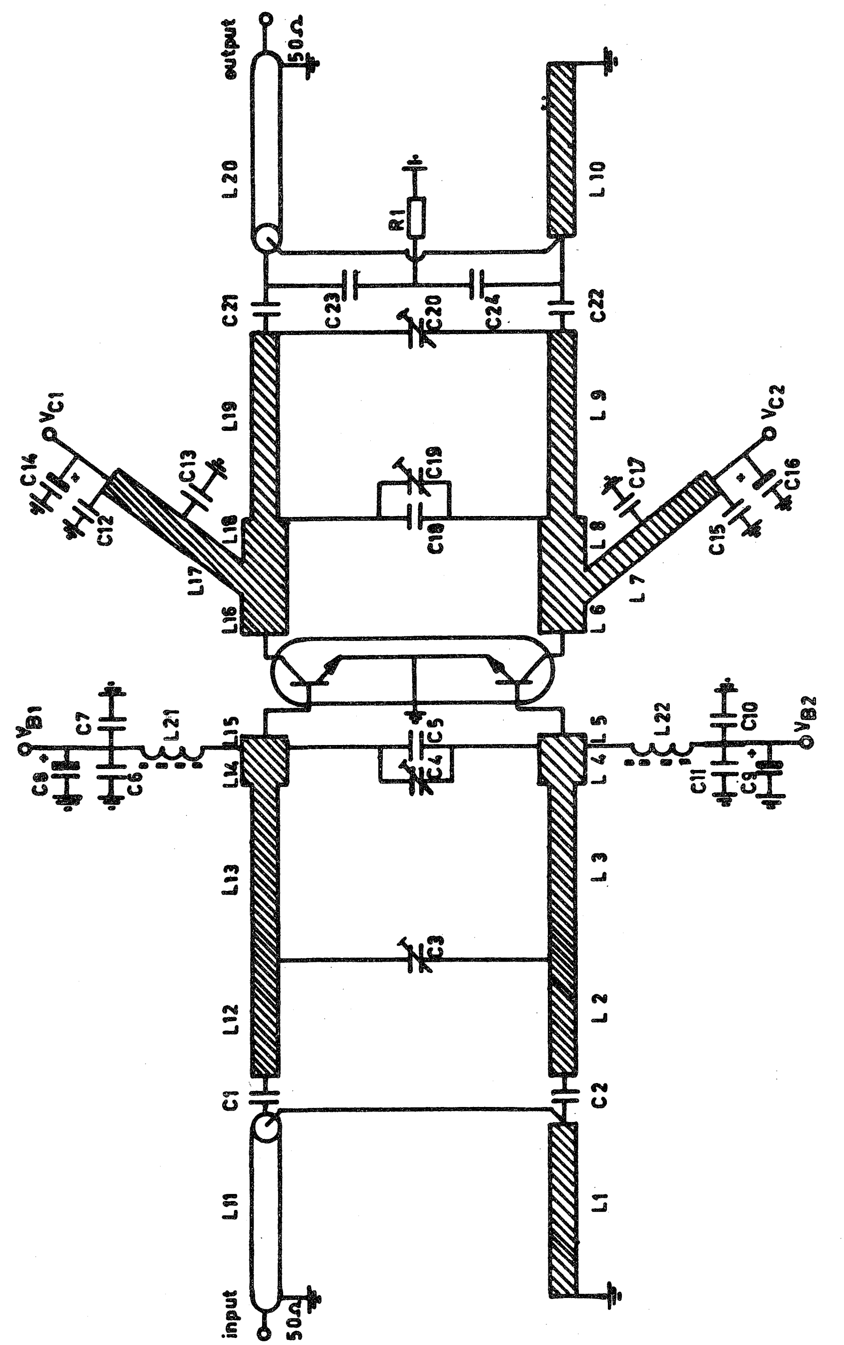

Fig.4 Circuit of one BLV 57 branch

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PHILIPS

6. LIST OF COMPONENTS BLV 57 CLASS AB (ONE BRANCH)

$C_1=C_2$=12pF chip capacitor, Philips NPO, cat.no.2222 851 13129

$C_3=C_4=C_{19}=C_{20}$=1-3.5pF film dielectric trimmer, Philips cat.no.2222809 05001

$C_5=C_{18}$=8.2pF chip capacitor, ATC, 8R2J

$C_6=C_{11}=C_{12}=C_{15}$=100nF chip capacitor, Philips NPO, cat.no.2222855 48104

$C_7=C_{10}$=100pF chip capacitor, Philips NPO, cat.no.2222852 13101

$C_8=C_9=C_{14}=C_{16}$=6.8µF,40V,electrolytic capacitor,Philips,cat.no.2222 030 37688

$C_{13}=C_{17}$=110pF chip capacitor, ATC, 111J

$C_{21}=C_{22}$=22pF chip capacitor, Philips NPO,cat.no.2222 851 13229

$C_{23}=C_{24}$=2.2pF chip capacitor, Johanson,no.500R,15N 2R2BA

L_1= stripline (Z_C=50Ω), 49x2mm

$L_2=L_{12}$=stripline (Z_C=57Ω), 14.5x1.5mm

$L_3=L_4$=stripline (Z_C=57Ω), 12.8x1.5mm

$L_4=L_{14}$=stripline (Z_C=36Ω), 2x3mm

$L_5=L_{15}$=stripline (Z_C=36Ω), 1x3mm

$L_6=L_{16}$=stripline (Z_C=36Ω), 3x3mm

$L_7=L_{17}$=stripline (Z_C=48Ω), 17.7x2mm

$L_8=L_{18}$=stripline (Z_C=36Ω), 8.8x3mm

$L_9=L_{19}$=stripline (Z_C=57Ω), 15.2x1.5mm

L_{10}= stripline (Z_C=50Ω), 46x2mm

L_{11}=49mm semi-rigid coax, 2.2mm Ø, Z_C=50Ω, PTFE dielectric, soldered on 2mm stripline

L_{20}=46mm semi-rigid coax, 2.2mm Ø, Z_C=50Ω , PTFE dielectric, soldered on 2mm stripline

$L_{21}=L_{22}$=0.1µH microchoke, cat.no.2322 057 01071

R=12Ω ,CR25 type, cat.no.2322 211 13129

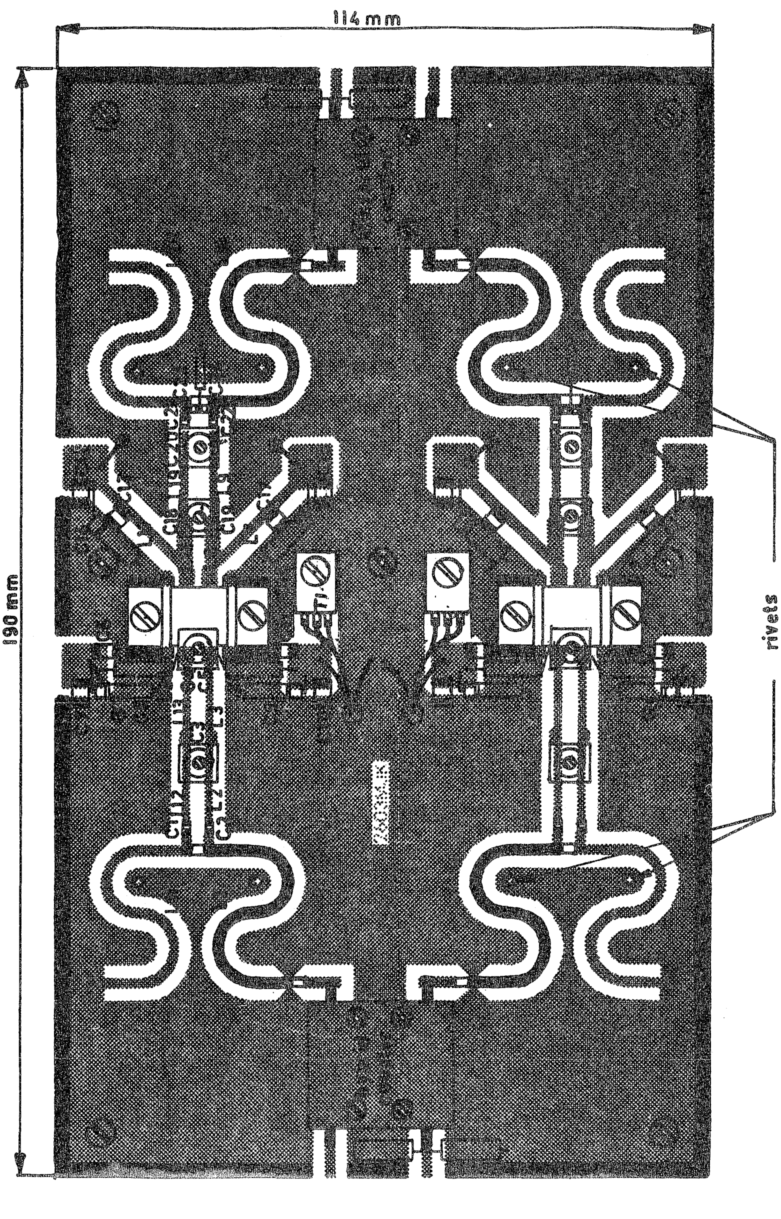


Fig.5 Lay-out of the BLV 57 amplifier

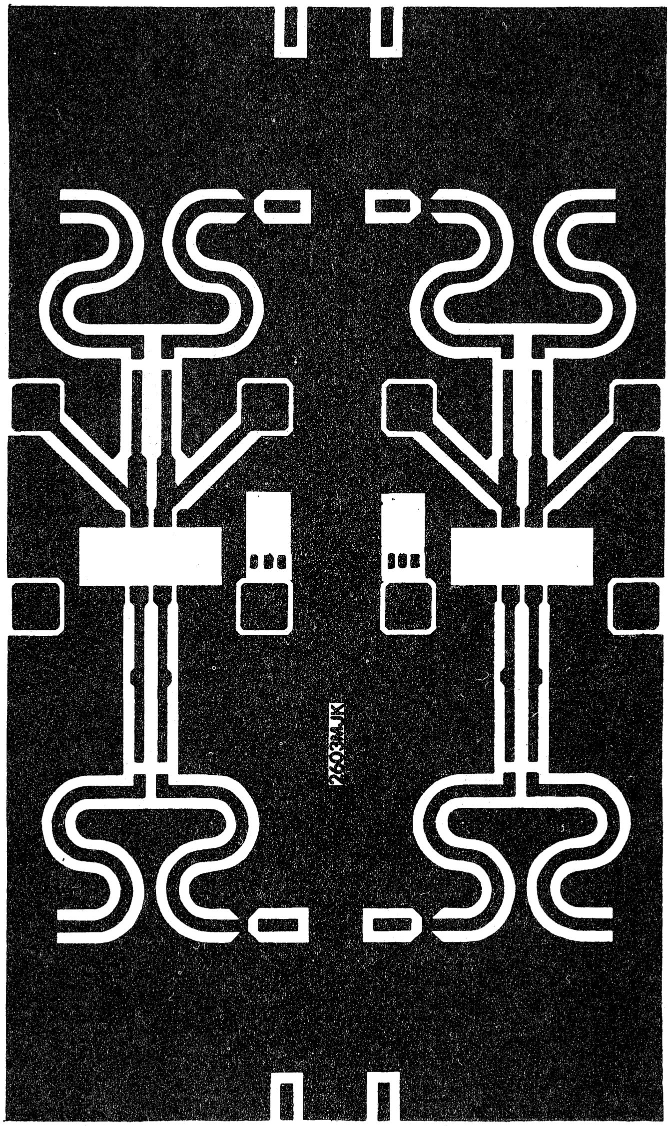


Fig.6 p.c.-board of the BLV 57 amplifier

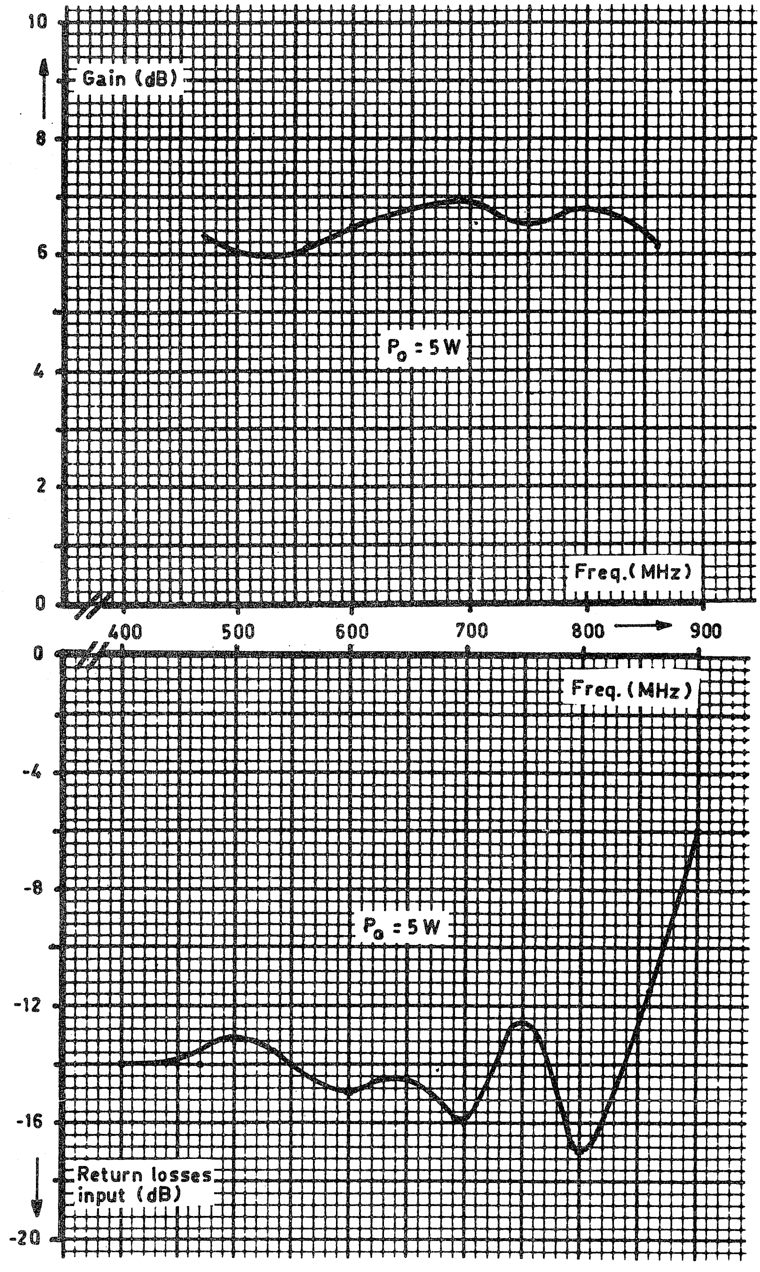
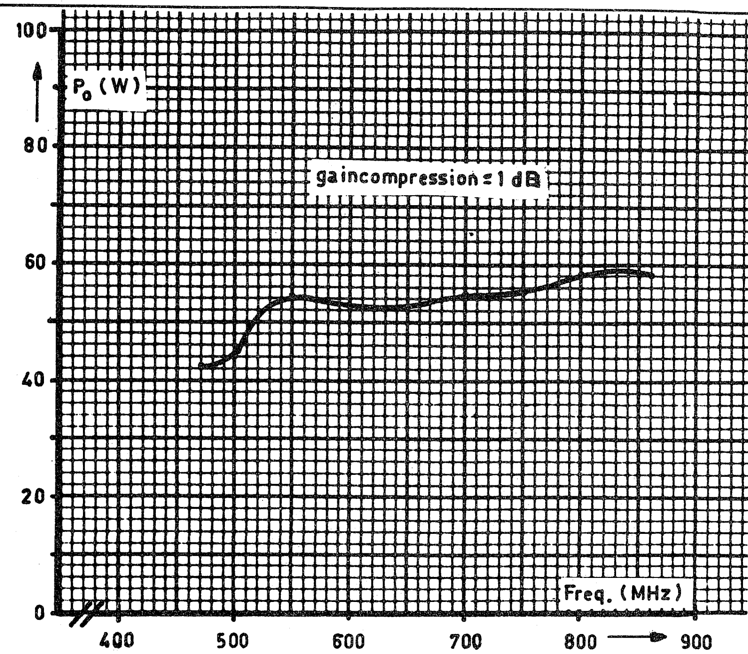


Fig.7 Gain and input return losses



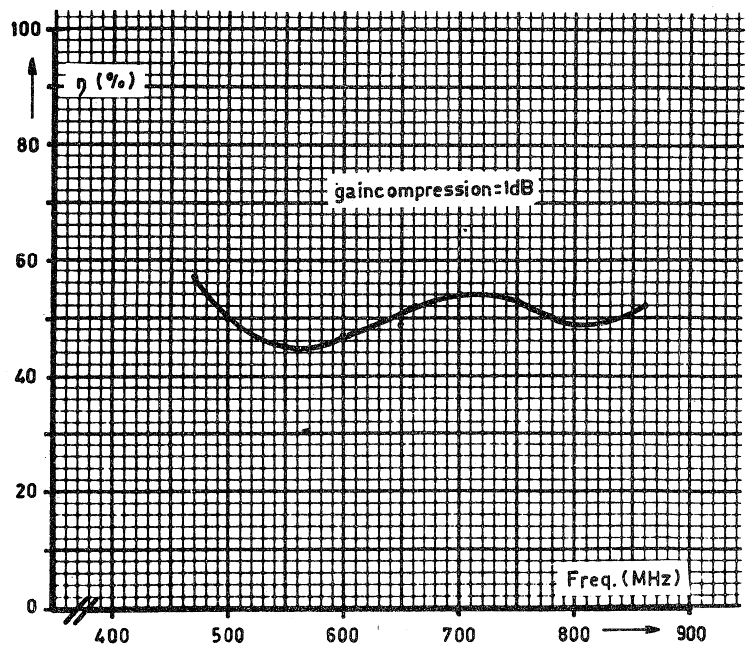


Fig.8 Gain versis output power

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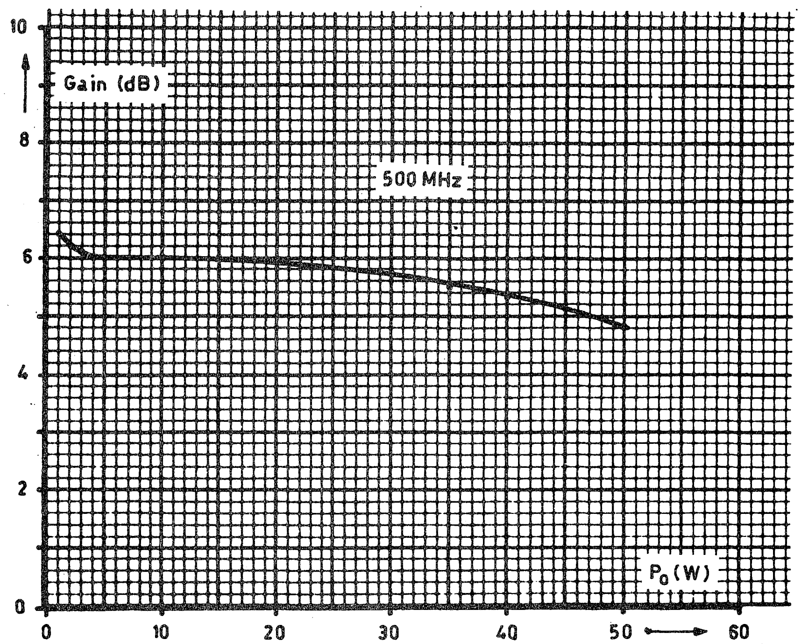

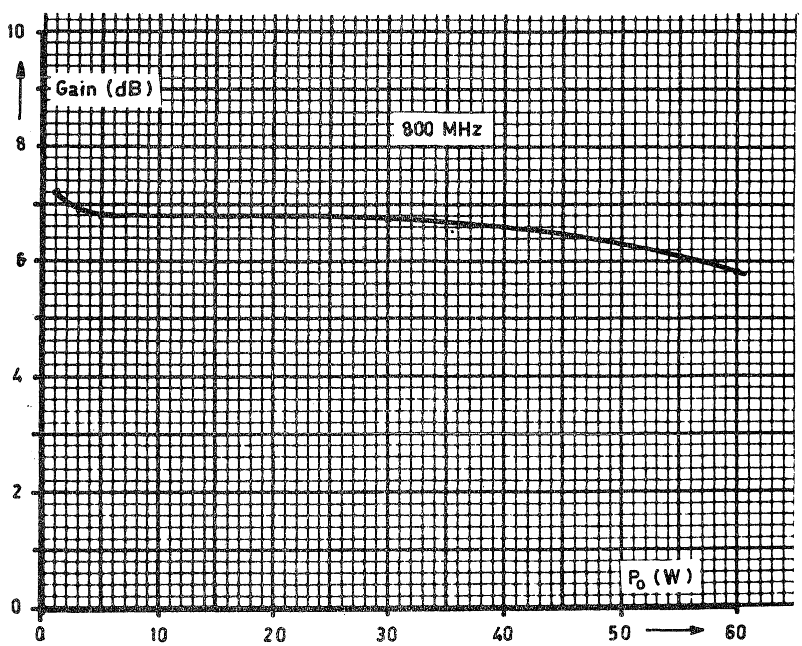

Fig.9 Output power and efficiency

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N.V. PHILIPS SEMICONDUCTORS NIJMEGEN – THE NETHERLANDS

PRODUCT GROUP SPECIALTIES AND DIODES

Report no. : RNR-1-233-1982-AS / NCO 8206		**APPLICATION**
Author : H.van Hees		
Date : 1982-10-12		

WIDEBAND CLASS AB POWER AMPLIFIER FOR TV TRANSPOSERS
AND TRANSMITTERS IN BAND I (50-80MHz) WITH TWO TRANSISTORS
BLV 36

SUMMARY

The transistor BLV 36 is primarily intended for use in linear VHF amplifiers for television transposers and transmitters. In report NCO 8103 Mr.v.Hees describes a wideband class AB amplifier with the BLV 36 for the TV band III (174-230 MHz).

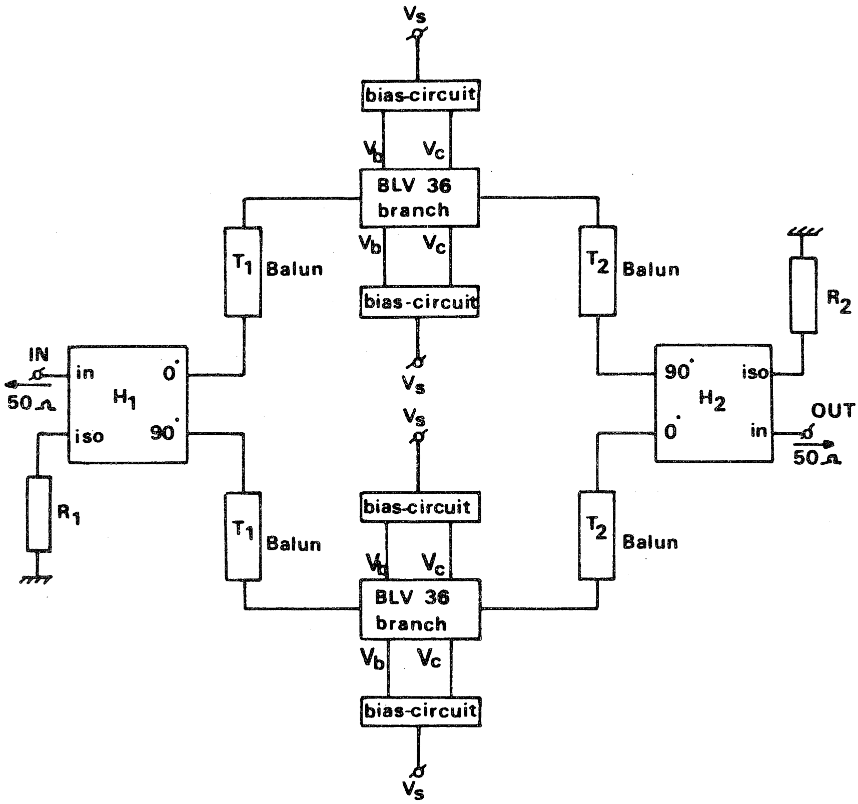
On customers request we have also made a theoretical design of such an amplifier for the TV band I (50-80MHz).

The expected performancc is:

Frequency range	:	band I 50-80MHz
Gain at P_{out}=200W	:	17.2dB $\pm$ 0.5dB
P_{out} at 1dB gain compression	:	$\geq$ 250 Watt
Gain at 1dB gain compression	:	$\geq$ 16dB
Input return losses	:	$\geq$ 18dB
D.C.setting BLV 36	:	I_{cz} = 2x150mA
		V_{CE} = 28 Volt

Fig.1 shows the schematic line-up of the complete amplifier.
The applied bias-circuit has been described in the appendix of report NCO 8103.
The construction of the baluns T_1 and T_2 is illustrated in Figs.3 and 4.
Fig.2 gives the circuit of one BLV 36 branche

H.van Hees



$H_1=H_2$ = 90°-3dB hybrid coupler.
type Anaren 1 A0230-3.

$R_1=R_2$ = 50Ω resistor.
type Pyrofilm PPR 515-20-3

T_1 and T_2 see Fig.3 and Fig.4

Fig.1. Schematic line-up of the amplifier.

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PHILIPS

Electronic components and materials

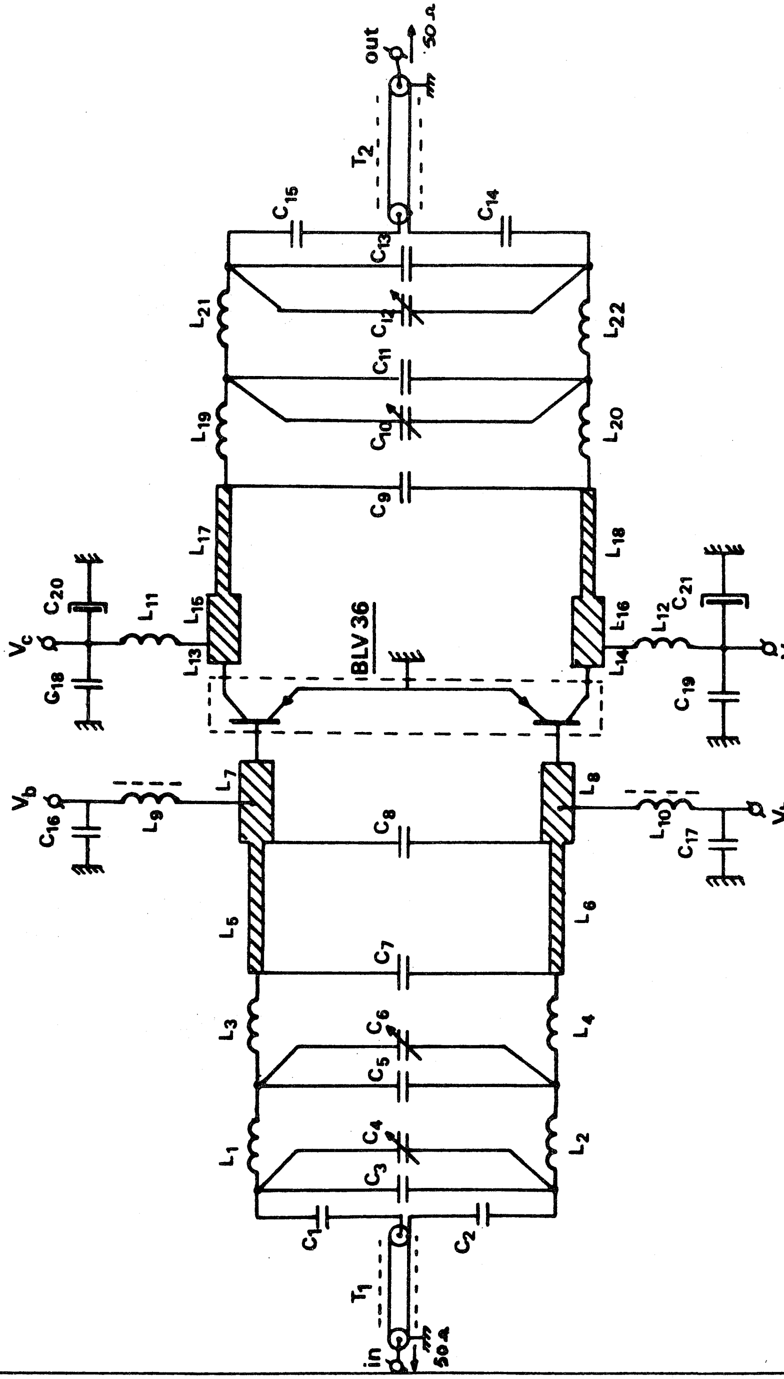

Fig.2. Circuit of a BLV 36 branch.

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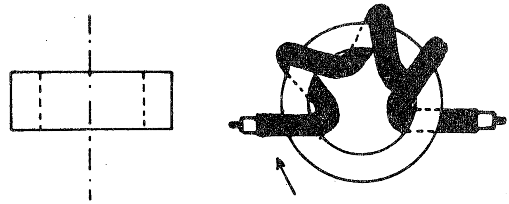
4C6 toroid

14x9x5 mm

Philips cat.no. 4322 020 91020



50Ω semi-rigid coaxial cable.

type Suhner RG 178 B/U.

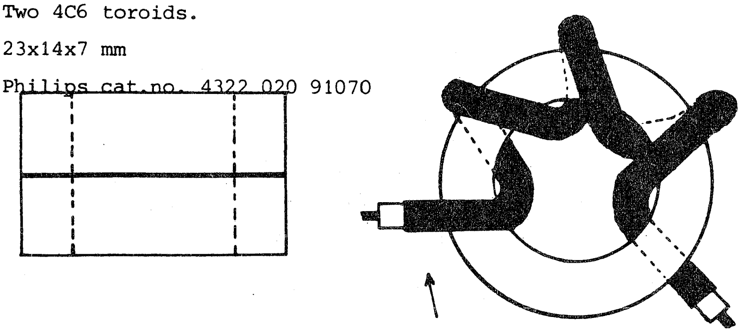
d_{out} = 1.8mm

Fig.3. Balun T_1. Scale 2:1.

Two 4C6 toroids.

23x14x7 mm

Philips cat.no. 4322 020 91070



50Ω semi-rigid coaxial cable.

type Suhner RG 188 A/U.

d_{out} = 2.6mm

Fig.4. Balun T_2. Scale 2:1

Electronic components and materials

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PARTS LIST

$C_1=C_2=C_{15}=C_{14}=$ 2K7 chip capacitor Philips cat.no.2222 852 13272

$C_3=$27pF ATC 100B.

$C_4=C_6=C_{10}=C_{12}=$0-40pF trimmer Philips cat.no. 2222 809 07009.

$C_5=$110pF ATC 100B.

$C_7=$110pF ATC 100B.

$C_8=$680pF ATC 100B.

$C_9=$240pF ATC 100B.

$C_{11}=$130pF ATC 100B.

$C_{13}=$30pF ATC 100B.

$C_{16}=C_{17}=C_{18}=C_{19}=$2K7 chip capacitor Philips cat.no.2222 852 13272.

$C_{20}=C_{21}=$47µF/40V elco Philips cat.no. 2222 030 27479.

$L_1=L_2=$52.6nH. 4 turns enamelled Cu-wire Ø 1mm, int diam.D=4mm, length 5.9mm, leads 2x5mm.

$L_3=L_4=$29.4nH. 4 turns enamelled Cu-wire, Ø 1mm,int.diam.D=3mm, length 8.1mm, leads 2x5mm.

$L_5=L_6=$41.2Ω stripline l=39.8mm, W=2mm.

$L_7=L_8=$31.2Ω stripline l=11.3mm, W=3mm.

$L_9=L_{10}=$100nH µ-choke, Philips cat.no. 4322 057 01071.

$L_{11}=L_{12}=$50nH 4 turns enamelled Cu-wire, Ø 1.5mm, int.diam.D=5mm, length 11.3mm, leads 2x5mm.

$L_{13}=L_{14}=$31.2Ω stripline, length l=5mm, W=3mm.

$L_{15}=L_{16}=$31.2Ω stripline, length l=9.4mm,W=3mm.

$L_{19}=L_{20}=$27.6nH 3 turns enamelled Cu-wire, Ø 1.5mm, int.diam.D=4mm, length 8.5mm, leads 2x5mm.

$L_{21}=L_{22}=$47.3nH 4 turns enamelled Cu-wire, Ø 1.5mm, int.diam.D=5mm, length 12.2mm, leads 2x5mm.

$T_1=$balun 4 turns coaxial cable type RG 178 B/U, on a 4C6 toroid (14x9x5mm cat.no. 4322 020 91020.)

$T_2=$balun 4 turns coaxial cable type RG 188 A/U, on two 4C6 toroids (23x14x7mm, cat.no.4322 020 91070).

p.c.board material is epoxy fibre-glass ($E_r \approx 4.5$), thickness 0,8mm.

N.V. PHILIPS SEMICONDUCTORS NIJMEGEN – THE NETHERLANDS

PRODUCT GROUP SPECIALTIES AND DIODES

Report no. : RNR-1-234-1982-AS / NCO 8207 **Author** : H.van Hees **Date** : 1982-10-12	**APPLICATION**

WIDEBAND CLASS A POWER AMPLIFIER FOR TV TRANSPOSERS IN BAND I (50-80MHz) WITH TWO TRANSISTORS BLV 33

SUMMARY

The transistor BLV 33 is primarily intended for use in linear VHF amplifiers for television transmitters and transposers. In report ECO-7904 Mr.Zwanen describes the application of the BLV 33 in a transposer amplifier for the TV band III (174-230MHz).

On customers request we have made a theoretical design of such a linear transposer amplifier for the TV band I (50-80MHz).

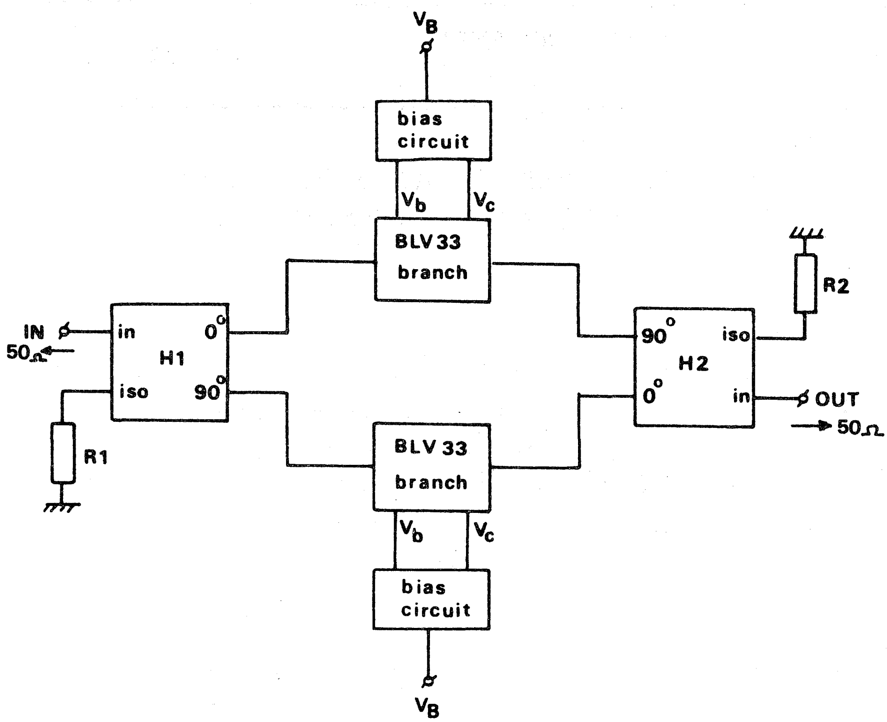
Expected performance:

Frequency range	band I 50-80MHz
Gain	17.4dB $\pm$ 0.3dB
Input VSWR	$\leq$1.25
Output VSWR	$\leq$1.25
Output power Po_{sync}	
at -60dB int	$\leq$30 Watt
at -55dB int	$\leq$45 Watt
Cross modulation	
at Po_{sync}=40Watt	$\leq$7.5%
D.C.setting BLV 33	I_c=3.2A/V_{CE}=25 Volt.

Fig.1 shows the schematic line-up of the complete amplifier and Fig.2 gives the circuit of one BLV 33 branch.

The applied bias-unit has been described in report ECO 7904.

H.van Hees



$H_1 = H_2 =$ hybrid coupler

Anaren type 1A0230-3

$R_1 = R_2 =$ 50Ω resistor

type Pyrofilm PPR 515-20-3

Fig.1. Schematic line-up of the amplifier.

PHILIPS

Electronic components and materials

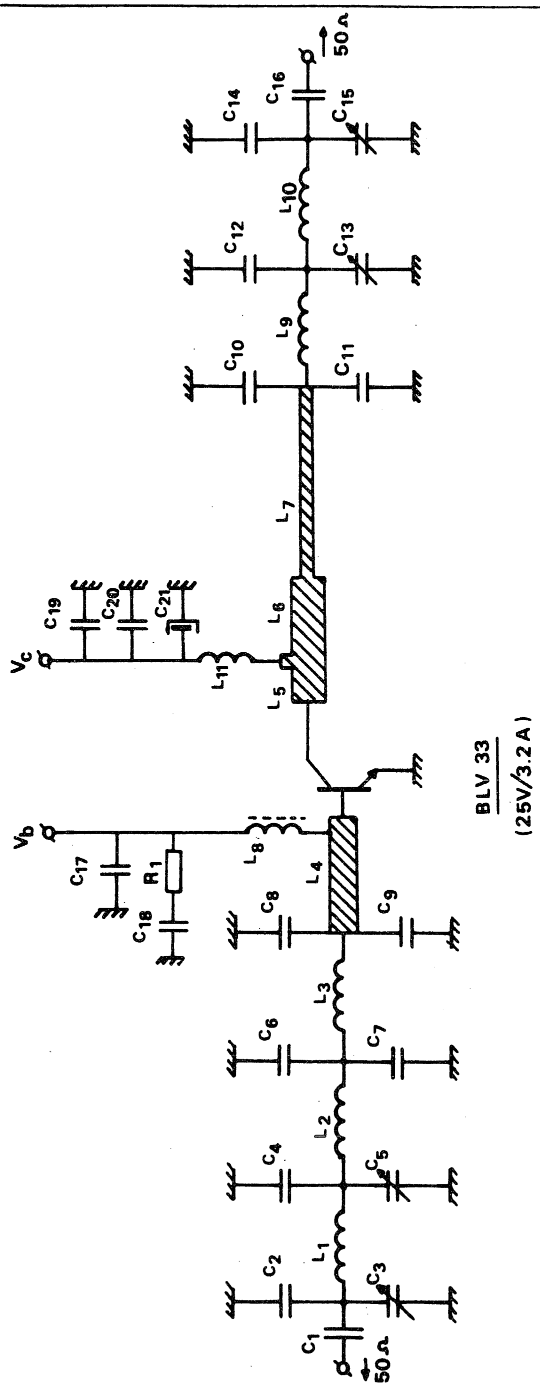


Fig.2. Circuit of a BLV 33 branch.

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PARTS LIST

$C_1=C_{16}=C_{17}=C_{19}$=2K7 chip capacitor Philips cat.no. 2222 852 13272.

$C_2=C_{14}$=39pF chip capacitor ATC 100B.

$C_3=C_5$=0-18pF film dielectric trimmer Philips cat.no. 2222 809 05003.

C_4=110pF chip capacitor ATC 100B.

$C_6=C_7$=82pF chip capacitor ATC 100B

$C_8=C_9$=300pF chip capacitor ATC 100B

$C_{10}=C_{11}$=160pF chip capacitor ATC 100B.

C_{12}=150pF chip capacitor ATC 100B.

$C_{13}=C_{15}$=0-40pF film dielectric trimmer. Philips cat.no.2222 809 07009.

$C_{18}=C_{20}$=330nF polyester capacitor. Philips cat.no. 2222 352 25334.

C_{21}= 47µF elco Philips cat.no. 2222 030 37479

L_1=103 nH 5 turns enemalled Cu-wire, Ø 1mm, int.diam.D=5.5mm, length 7.6mm, leads 2x5mm

L_2=61.3nH 5 turns enamelled Cu-wire,Ø 1mm,int.diam.D=4.5mm,length 10.6mm, leads 2x5mm

L_3=27.1nH 3 turns enamelled Cu-wire,Ø 1mm,int.diam.D=4.4mm, leads 2x5mm.

L_4=30.1Ω stripline width 6mm, length l=30mm

L_5=30.1Ω stripline width W=6mm, length l=5mm

L_6=30.1Ω stripline width W=6mm, length l=10.6mm

L_7=60.2Ω stripline width W=2mm, length l=35.5mm

L_8=1µH choke Philips cat.no.4322 057 01081.

L_9=43.8nH 4 turns enamelled Cu-wire, Ø 1.5mm, int.diam.D=4.5mm,length 11.3mm, leads 2x5mm.

L_{10}=87.4nH 5 turns enamelled Cu-wire, Ø 1.5mm,int.diam.D=6.5mm,length 15.1mm, leads 2x5mm

$L_{11}=L_{10}$

R_1=10Ω metal film resistor.

p.c. board material is epoxy fibre-glass (E_r=4.5), thickness 1/16".

N.V. PHILIPS SEMICONDUCTORS NIJMEGEN – THE NETHERLANDS

PRODUCT GROUP SPECIALTIES AND DIODES

Report no.: RNR-1-308-1982-AS / NCO 8208 **Author** : H.v.Hees **Date** : 1982-12-13	**APPLICATION**

WIDEBAND CLASS A POWER AMPLIFIER FOR TV TRANSPOSERS IN BAND I (54-88MHz) WITH TWO TRANSISTORS BLV 31

SUMMARY

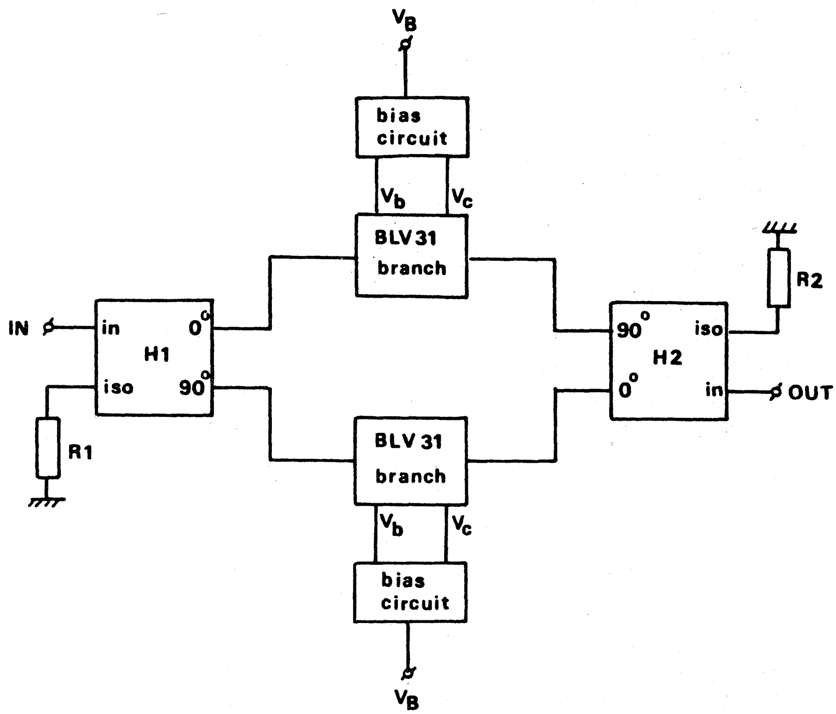
The transistor BLV 31 is primarily intended for use in linear VHF amplifiers for television transmitters and transposers.
In report ECO 8003 Mr.Zwanen/Boekhoudt describe the application of the BLV 31 in a transposer amplifier for the TV band III (174-230MHz).
On customers request we have made a theoretical design of such a linear transposer amplifier for the TV band I (54-88MHz).

EXPECTED PERFORMANCE

Frequency range	band I 54-88MHz
Gain	21dB $\pm$ 0.3 dB
Input VSWR	$\leq$1.25
Output VSWR	$\leq$1.25
Output power Po_{sync}	
at -60dB int	10 Watt
at -57dB int	12 Watt
Cross modulation	
at Po_{sync}=10W	$\leq$8%
D.C.setting BLV 31	I_C=0.8A/V_{CE}=25V

Fig.1 shows the schematic line-up of the complete amplifier and Fig.2 gives the circuit of one BLV 31 branch.
The applied bias-unit has been described in report ECO 8003.

H.van Hees



$H_1=H_2=$ hybrid coupler
Anaren type 1 A0230-3

$R_1=R_2=$ 50Ω resistor
type Pyrofilm PPR 515-20-3

Fig.1 Schematic line-up of the amplifier.

Electronic components and materials

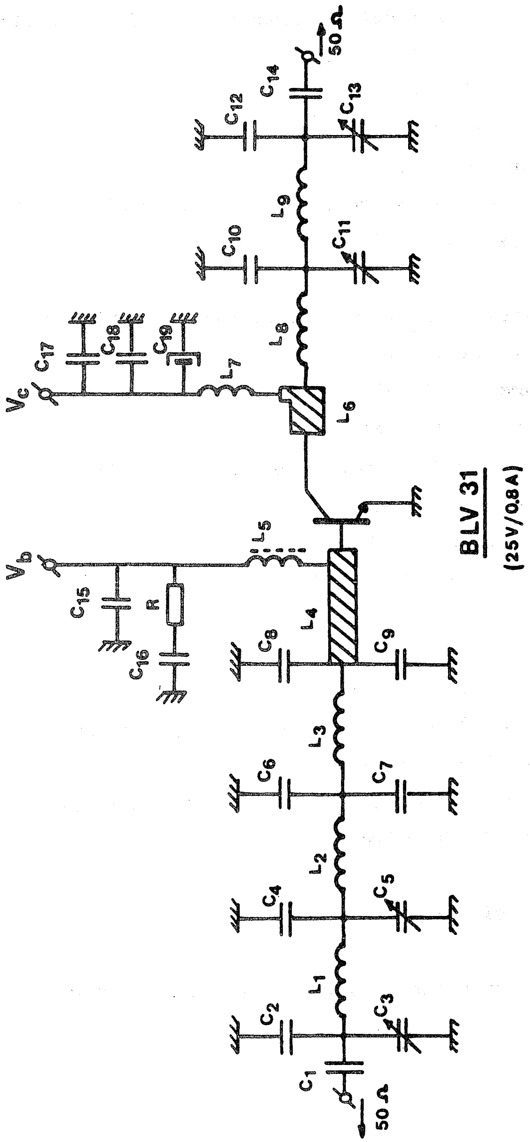


Fig.2 Circuit of a BLV 31 branch

Semiconductor application laboratory SAL Nijmegen - The Netherlands

PARTS LIST

$C_1=C_{14}=C_{15}=C_{18}$=2K7 chip capacitor Philips cat.no. 2222 852 13272

$C_2=C_{12}$=30pF chip capacitor ATC 100B

$C_3=C_5=C_{11}=C_{13}$=2-18pF film dielectric trimmer cat.no. 2222 809 05003

C_4=82pF chip capacitor ATC 100B.

$C_6=C_7$=68pF chip capacitor ATC 100B

$C_8=C_9$=200pF chip capacitor ATC 100B

C_{10}=75pF chip capacitor ATC 100B

$C_{16}=C_{17}$=330nF polyester capacitor Philips cat.no. 2222 352 25334

C_{19}=47µF elco Philips cat.no. 2222 030 37479

L_1=101 nH 5 turns enamelled C_u-wire, Ø 1mm, int.diam.D=5.5mm
length 7.8mm, leads 2x5mm

L_2=67.5 nH 5 turns enamelled Cu-wire, Ø 1mm, int.diam.D=4mm,
length 7.5mm, leads 2x5mm

L_3=25.3nH 3 turns enamelled Cu-wire, Ø 1mm, int.diam. D=4mm,
length 8.2mm, leads 2x5mm.

L_4=30.1 Ω stripline, width 6mm, length 39mm.

L_5=1µH choke Philips cat.no. 4322 057 01081

L_6=30.1Ω stripline, width 6mm, length 5mm.

L_7=131nH 6 turns enamelled Cu-wire, Ø 1mm, int.diam.D=6mm
length 10.6mm, leads 2x5mm.

L_8=45.5nH 4 turns enamelled Cu-wire, Ø 1mm, int.diam.D=4mm
length 7.3mm, leads 2x5mm

L_9=93nH 5 turns enamelled Cu-wire, Ø 1mm, int.diam D=5.5mm
length 8.8mm, leads 2x5mm

R=10Ω metal film resistor

p.c. board material is epoxy fibre-glass ($\varepsilon_r \simeq 4.5$), thickness 1/16".

PHILIPS | PRODUCT GROUP **SPECIALTIES AND DIODES** NIJMEGEN | APPLICATION

Report no : RNR45/072/1988/AS / NCO 8803
Author : M.J.Köppen
Date : 1988-04-07

A WIDEBAND SINGLE STAGE LINEAR POWER AMPLIFIER EQUIPPED WITH 2X BLV38 FOR BAND III T.V.

SUMMARY

In this report a description is given of a wideband power amplifier for television services in band III (174-230MHz).

The amplifier is intended for vision amplification and operates in class-AB.

It utilizes two push-pull transistors BLV38. These devices have internal prématching. Both transistors are coupled by means of 3dB-90^{o} coaxial hybrids.

The nominal output power amounts to around 450W for -1dB power gain compression at a supply voltage of 35V.

1. INTRODUCTION

For application in television band III (174-230MHz) a wideband power amplifier has been designed with two transistors BLV 38.

The BLV38 is a push-pull device i.e., two identical transistor chips are mounted in a single case and driven 180^{o} out of phase.

This case is a 5-lead envelope with ceramic cap. (SOT-179).

The BLV38 has been developed to operate in class-AB with zero signal collector idling currents of approx. 2x200mA from a stabilized collector supply voltage of 35V.

For optimal operation and matching the BLV38 is pré-matched.

This report describes the design, construction and practical results of the amplifier.

2. DESIGN OF THE AMPLIFIER

Fig.1 shows the block diagram of the complete amplifier.
Essential elements are:
- The single BLV38 wide-band amplifier
- The coaxial 3dB-90^{o} hybrid
- The bias circuit.

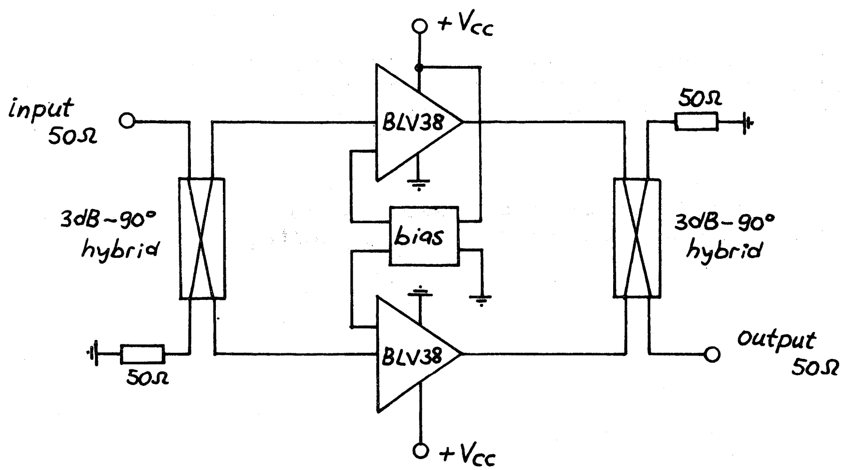


Fig.1.

These three elements will be discussed in this order.

3. THE SINGLE BLV38 WIDE-BAND AMPLIFIER

3.1. Properties of the transistor and some info of the test method.

The Data Handbook publication is as follows:

f = 224.25MHz; V_{CE}=35V; $I_{C(ZS)}$=2x0.2A; T_h=25°C
P_L =225W ; G_p>8.0dB;G_p typical 8.8dB
η_c>50% ;η_c typical 58%
P_L =112.5W ; G_p>9.0dB;G_p typical 9.8dB

dG_p at 225W $\leqslant$1.0dB; typical 0.65dB.

These values are valid in a wide-band circuit being tuned at the highest frequency 224.25MHz.
Further:
Assuming a 3rd order amplitude characteristic, 1dB gain compression corresponds with 30% sync input/25% sync output compression in television service (negative modulation, CCIR system).

The, here described, amplifier has been created during the final phase of development of the BLV38. In that period the transistor design was ready, but exact impedances as a function of frequency were unknown. Because the author has to do both viz. developing the BLV38 and performing applications it has been started with the practical measurement of the input- and load impedances of the device at three spot frequencies 175, 200 and 225MHz.

Up to that moment the only available circuit was the semi-wideband test circuit. This circuit has been designed for transistor impedances being calculated with the aid of the equivalent circuit diagram of the BLV38.

Except for a somewhat uncertain part of this equivalent diagram viz. the mutual coupling between the bonding-wires of base and emitter in the pré-matching section, rather reliable complex impedance figures could be predicted. Surely, in the case of the relative low frequency range up to 225MHz.

In principle these calculations are done on the single-ended transistor, or in this case one half BLV38. The earth line then is equal to the virtual one inside the device.
A practical impedance check is advisable. However in that case one needs a circuit that can be tuned from at least 175 to 225MHz and handle the rather high power.

This specific test circuit has been constructed for 50Ω input and output impedance to earth and is the one being published in our Data Handbook.

Although, as has been stated before, the latter circuit in principle has been developed for wideband operation we changed it slightly i.e. the tuning range is somewhat more accented to the highest frequency 225MHz. So, some improvisations were needed.

3.2. Impedance measurements

Although, it is quite easy only to repeat here the already published impedances of the Data Handbook, some info will be given about the measuring and calculation methods.

A possible method to measure the actual complex impedances belonging to a matched transistor for a given frequency, power, supply voltage, class of operation and temperature is to make first the device operational under these conditions and measure then the passive properties of the networks around the transistor.

However these networks can be rather complicated. For example in case of the BLV38 they contain a three-stage impedance transformation and a shortened balun.

So, the well known method of removing the transistor, matching the input and output with the characteristic impedance (50Ω) and remeasuring the conjungate or direct complex impedance into the matching sections does not work here very well because of the following handicaps:
- the input and output are balanced
- the impedances are strongly deviating from 50Ω.

Another possibility is to measure from the 50Ω input or output into the direction of the removed transistor.

It can be proved that the actual push-pull transistor impedances can be calculated then from the complex reflection coefficients (50Ω level) occuring if one exchanges the transistor resp. by a short-circuited, a plain, and some with known, but different (3) resistors (and inductances) equipped, dummy encapsulations.

In this way, the already calculated impedances, have been checked. The correlation with the calculated impedances was acceptable (see chapter 3.5.).

3.3. <u>The output matching network</u>

Fig.2 shows the configuration of the output network.

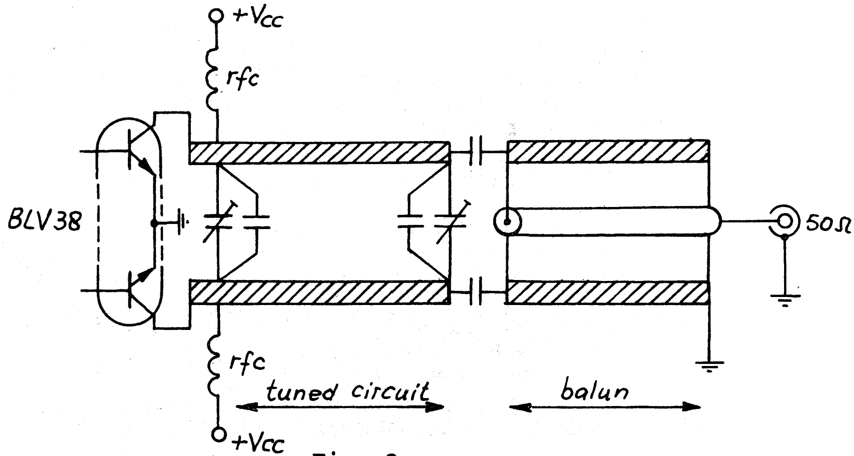


Fig. 2.

This network may be divided again in two parts viz. the tuneable part and the balun (balanced-unbalanced) transformer.

It has been drawn here according to the ultimate diagram, however for calculations it is much easier to apply the asymmetrical model (Fig.3) including the parasitic series inductances of the capacitors.

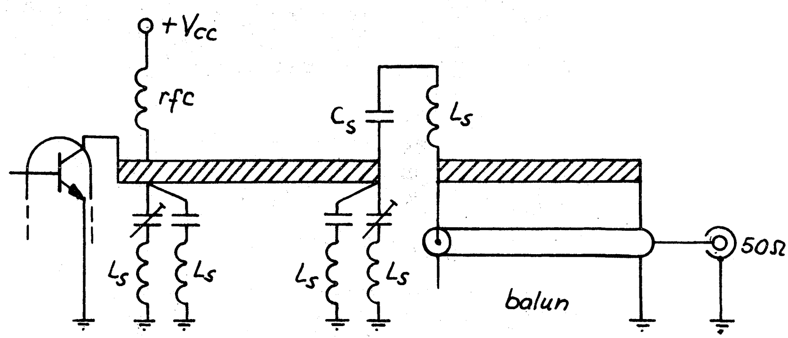


Fig. 3.

3.4. Value of RFC

The collector choke should have three functions:
1) supplying the d.c. to the collector
2) making part of the transformation
3) preventing parasitic oscillations at low frequencies (MHz range).

1) Is necessary, 2) may help sometimes, but is not advisable here, 3) is a must.
An advise value for the RFC can be calculated from
$X_L \approx 5.Rp$ and
$R_p \approx (V_{CE}-V_{CE\ sat})^2/\ 2.P_L\Omega$
So $L(nH) \approx 800\ (V_{CE}-V_{CE\ sat})^2/f.P_L=34.42nH$ (f in MHz)
It has been chosen for 30nH.

3.5. Published output impedance versus frequency

f (MHz)	$R_L \pm jX_L (\Omega)$
150	2.86 +j 1.61
170	2.57 +j 1.54
190	2.30 +j 1.44
210	2.06 +j 1.31
230	1.85 +j 1.16
250	1.66 +j 1.01

Conjugate complex diagram

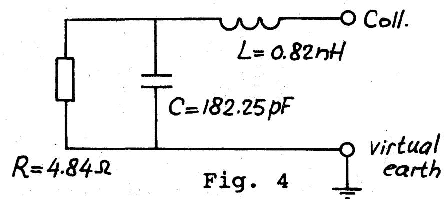


Fig. 4

As has been mentioned before the impedances are calculated and checked under practical conditions.
The published data are calculated from the quite simple equivalent diagram of Fig.4 and so somewhat streamlined.
In this case:
$R = (V_{CE}-V_{CEsat})^2/2.P_L=(35-2)^2/2.112.5 = 2.7\Omega$
$C = 1.15\ (C_{cbi} + C_{cbo} + C_{ce}) + C_s = 182.25pF$
$L = L_{b/2} + L_c \approx 0.82nH$

The $R_L \pm jX_L$ are the conjungate complex impedances.

Our practical check gave:
$f = 175MHz;\ R_L \pm jX_L = 2.17 + j\ 1.65\Omega$
$f = 200MHz;\ R_L \pm jX_L = 2.03 + j\ 1.43\Omega$
$f = 225MHz;\ R_L \pm jX_L = 1.56 + j\ 1.33\Omega$
So rather reliable figures. These values were used for the designs of the here described amplifier. With the aid of a computer programme it is possible again to find the best fitting 3-components network like Fig.4. The elements with an maximum error of 5.1% have been calculated to be R=5.19Ω; C=200.52pF and L=0.815nH.

For a frequency of 200MHz the recalculated impedance changes somewhat to 1.92 + j 1.48Ω in series-components or 3.06 and j 3.97Ω expressed in parallel components.

If one starts from publication values, the end-effect will be about the same but not exactly equal to, for example, the lengths of all transmission lines in the components list of the here described amplifier.

3.6. The balun transformer

Because one side, the transistor, is known it is advisable to look now to the balun section. The complete unit has been redrawn in Fig.5.

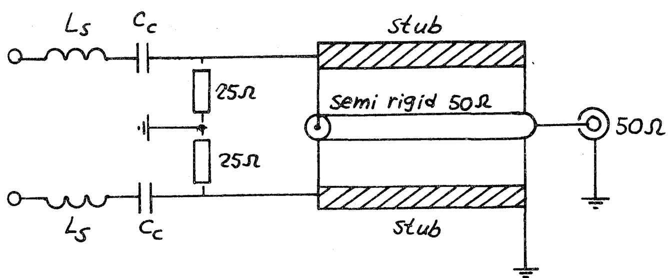


Fig. 5

Here it has been chosen for the situation that each transistor half of the BLV38 should transform via L_S and C_S to 25Ω. Doing so the influence of the semi-rigid coaxial cable can be neglected in first instance. The only elements playing a role in the transformation are then the stubs and the coupling capacitors C_C including their parasitic series inductances L_S.

In our design a physical length of 80mm has been applied. For Z_C=50Ω; ϵ_r=2.2; the short-circuited stub behaves as a function of frequency:
f = 175MHz; j 21.19Ω; 19.27nH
f = 200Mhz; j 24.66Ω; 19.62nH
f = 225MHz; j 28.32Ω; 20.03nH

or almost 20nH in parallel with 25Ω.

In mid-band (200MHz) the following transformation results:

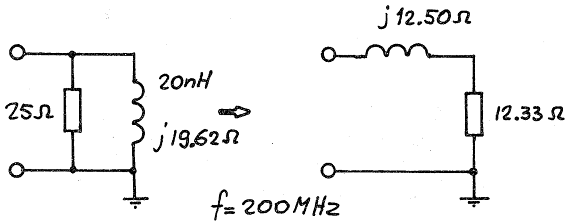


Fig. 6.

The coupling capacitor including parasitic series inductance can be given two functions i.e. firstly d.c. blocking and secondly impedance transformation.

It has to handle large r.f. currents so a parallel connection of more capacitors is necessary.

If one choose for $-jX_C + jX_{LS} = -j\ 12.5\Omega$ and $L_S \approx 0.5nH$ it can be done with three ATC100B capacitors of 18pF in parallel. This value slightly differs from the 60.61pF being asked for in mid-band (including L_S), so the resultant becomes 12.33-j 1.61Ω in series at 200MHz.

It should be noted here, that afore transformation components only have been taken at the mid frequency of 200Mhz. That is permitted here because all elements scale with the frequency and the wide-band performance will be optimized later-on with the aid of an optimizing computer programme.

3.7. <u>The transformation between single-ended BLV38 and balun transformer.</u>
Fig.7 shows the remaining situation between BLV38 and the balun.

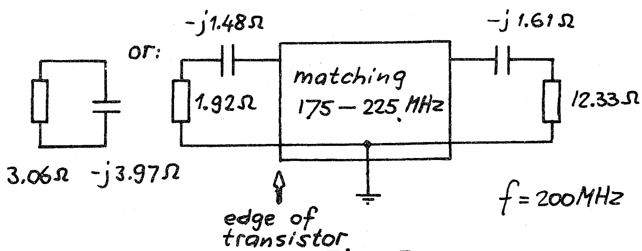


Fig. 7

The left hand part of Fig.7 represents the 200MHz situation at the edge (reference plane) of the transistor. To adapt the collector and the emitter lead some room is needed. The actual 1:1 situation is sketched in Fig.8.

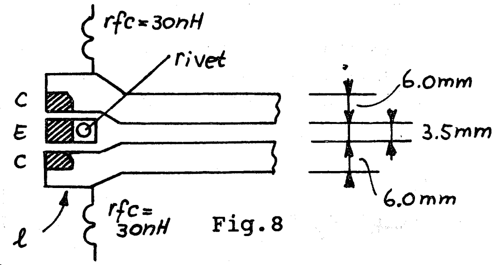


Fig.8

Material parameters are:
width stripline = 6.0mm
ϵ_r p.c. board = 2.2
d sheet = 2x0.035mm
D p.c. board = 1.59mm (1/16")

Stripline calculations learn:
Z_{Cdiel} = 42.86Ω
Z_{Cair} = 58.99Ω
$\sqrt{\epsilon_{eff}}$ = 1.38
The strip with length l=10mm can be considered as a stripline or a pi-section. For a pi-section one can calculate that it consists of the elements shown in Fig.9.

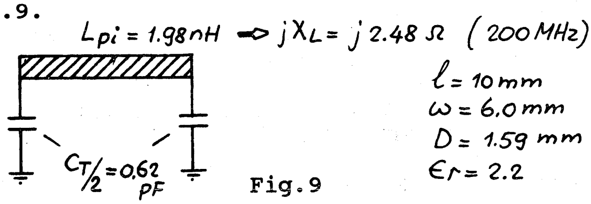


Fig.9

The influence of $C_T/2$ has been neglected, so up to the RFC=30nH the impedance amounts to: 1.92 + j 1.0Ω. Inserting the 30nH the impedance changes to 1.82 + j1.06Ω. So Fig.7 moves into Fig.10

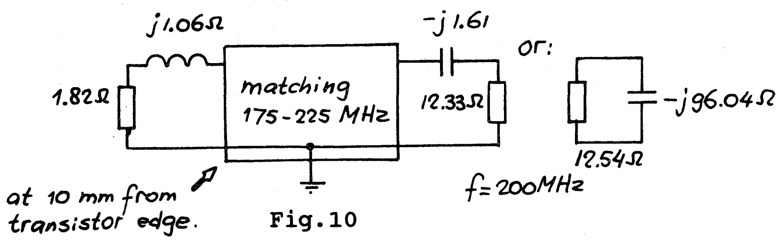


Fig.10

The matching section can be composed so that it shows a Chebishev response. Between 175 and 225MHz the following transformation (1.82 to 12.54Ω) can be done with a software programme:

nr.	VSWR	L1(nH)	C2(pF)	L3(nH)	C4(pF)	L5(nH)	C6(pF)
1	1.726	3.488	152.823				
2	1.073	2.083	281.046	6.414	91.270		
3	1.009	1.431	351.181	4.429	194.065	8.015	62.703

The two section solution has been taken with a maximum (calculated) VSWR of 1.073. Taking all transforming elements in one view Fig.10 will change in:

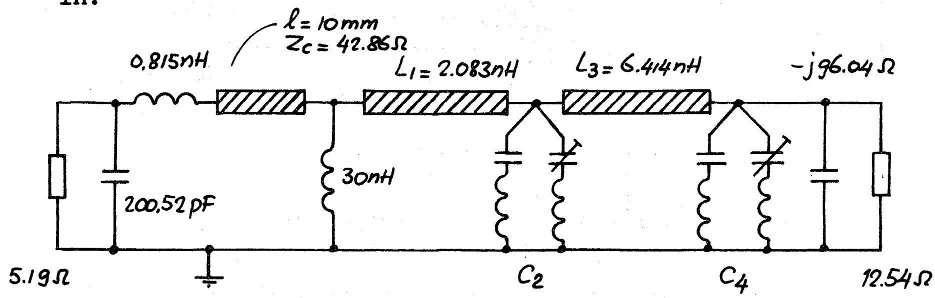


Fig.11

If the width and so Z_C of L_1 and L_3 are resp. equal to 6mm and 42.86Ω the C_T and L_{pi} are known. For D=1.59mm, ϵ_r=2.2 and $\sqrt{\epsilon_{eff}}$=1.38 they are:

C_T=1.235pF/10mm and L_{pi}=1.977nH/10mm.

This means that:

l_{el} of L_1=1.38.10.2.083/1.977=14.52mm and

l_{el} of L_3=1.38.10.6.414/1.977=44.74mm.

The capacitive parts consist of some fixed capacitors in parallel with a variable one.

According to experiences it has been chosen for (see Fig.12):

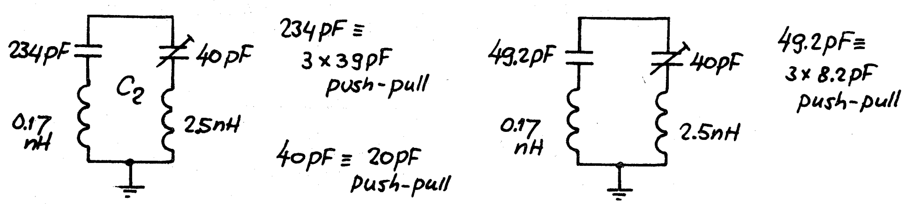


Fig.12

3.8. Optimization
All hand-calculated component values are put in a optimizing computer programme in which elements can be given free for optimization.
Some small corrections were needed, however the network is still close to the hand-calculated one.

3.9. The input matching network
In principle the design philosophy of this network is the same as has been applied for the output side. The balun (balanced-unbalanced) section is a direct copy.

As explained before, impedances have been simulated first and measured later on. Based upon experience the calculated impedances are liable to some insecurities. Because of the rather close correlation of the calculated and measured figures on the output stage, the measured values for the input of the BLV38 looked more reliable than the calculated ones, so the measured figures have been published in the Data Sheet.

4. **PUBLISHED INPUT IMPEDANCE VS. FREQUENCY**

For class-AB operation at V_{CE}=35V and P_L=225W these figures are as follows:

f(MHz)	Ri ± jXi (Ω)
170	0.50 + j 0.47
180	0.525 + j 0.57
190	0.54 + j 0.62
200	0.53 + j 0.64
210	0.51 + j 0.64
220	0.48 + j 0.63
230	0.45 + j 0.625

4.1. Some information of the input section
The design of this network shall not be described here in detail. Fig.13 shows the final configuration.

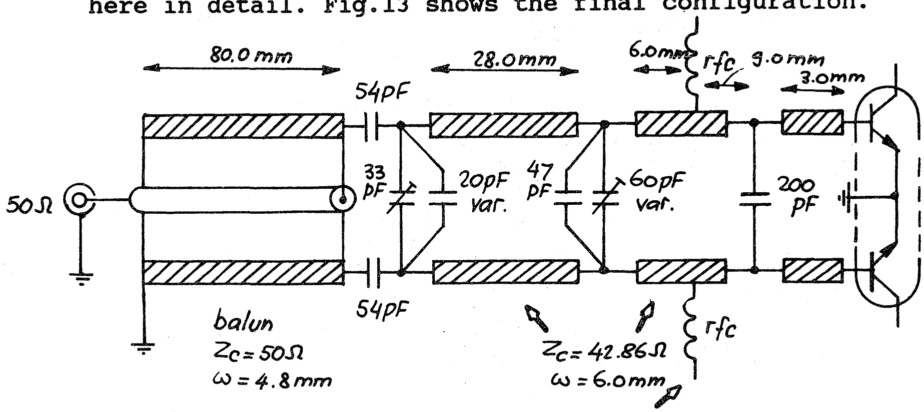


Fig.13

The capacitor of 200pF consists of 2x100pF in parallel. These components are soldered on top of each other, whilst the distance from the edge of the transistor cap to half the capacitors amounts to 3.0mm. The positioning of these capacitors strongly influences the tuning possibilities of the whole chain.

More info can be found in the complete circuit diagram (Fig.18) and the list of components.

5. THE COAXIAL 3dB-90°C QUADRATURE HYBRID

Experiments have been done with two configurations:
- the branch line type (Fig.14)
- 90°C crossover hybrid (Fig.15)

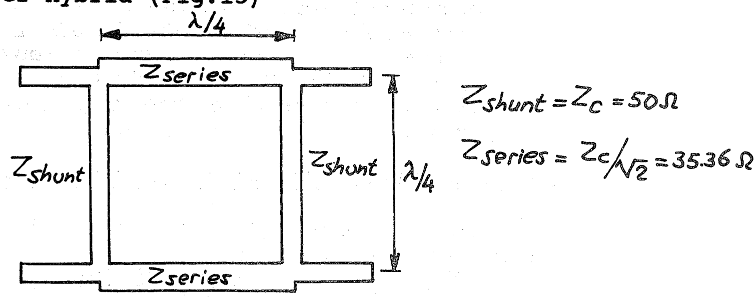


Fig.14

The branch line type is fundamentally a 90° hybrid; that is, the two output arms which are equal in amplitude are in a 90° phase relationship to each other. This 90° relationship is perfect only at the design frequency (f=200MHz) and varies with frequency. Although this 90° relationship is frequency sensitive, it varies only approx. ± 5° over a 10% bandwidth (here 20MHz).

Because the fundamental limitation is the bandwidth and we had no direct alternative available, first experiments have been done with two of these hybrids being printed on PTFE. The bandwidth restriction can be overcome by adding additional sections.

These units could handle 500W without problems.

Later on, it has been chosen for the second version the 90° crossover hybrid because it is smaller in size. Both power splitters/combiners have been built with pieces of SAGE WIREPAC type LCP1. Because the material may not be bent it has been chosen for a V-shape as sketched in Fig.15. The square semi-rigid coaxial lines are pressed to a brass plate on which each 4 N-type coaxial plugs are screwed.

Fig.16 shows the block diagram. Two equal units are needed.

Published isolation is greater than 30dB and a power rating of 500W up to 1 GHz is permitted.

6. THE BIAS SYSTEM

The BLV38 has to operate in class-AB. From some experiments it has been found that a suitable value for $I_{C(ZS)}$ amounts to approx. 200mA per section. Calculating with worst-case h_{FE}'s the bias circuit should be able to supply peaks of approx. 500mA.

Further it should have low internal resistance (not only for d.c.), temperature control and the possibility to balance both sections of one device.

Also it should be independent of supply voltage, short-circuit proof and able to withstand a break-through from collector to base of BLV38. The latter situation has to be indicated by an LED.

All these requirements are combined in the networks of Fig.18. To prevent l.f. detection of r.f. the unit is shielded and external connections are decoupled.
As shown in Fig.17 the temperature sensor transistors are mounted quite near to both BLV38's, so a firm temperature coupling is guaranteed. More details in Figs.17 and 18.

7. COMPLETE AMPLIFIER WITH TWO PIECES BLV38

Figs. 19 and 20 resp. show the complete circuit diagram and the components lay-out.

The amplifier consists of 3 main parts being screwed to an aluminium intermediate heatsink plate (256mmx136mmx10mm). Special attention has to be paid to the surface under the flanges of the transistors, it should be flat, clean and provided with a spare amount of heatsink grease.

For simplicity the p.c. boards for input-and output balun are the same. Size is 136mmx70mmx1/16". The semi-rigid cables are soldered to the stubs.

As Fig.20 shows the circuit and components are on one side of the PTFE fibre-glass board; the other side is unetched copper serving as ground plane.
Earth connections are made by hollow rivets and also by fixing screws and copper straps on input and output side of the board and under the emitters to provide a direct contact between the copper on the component side and the

It is advisable to press the perifery of both transistors down by means of some 2mm screws. Further copper foil has been applied between the earth side of the coaxial connectors and the lower sheet of the board. Also a sheet of copper foil is under the adjoining boards.

Because of losses the p.c. board should be RT/duroid 5880 of Rogers Corp. (1/16"; ϵ_r=2.2) or material with same quality.

The greater part of the fixed capacitors are ATC (American Technical Ceramics) capacitors type 100B or same quality.

The description of the p.c. boards has been done with the aid of a Circuit Mask (CMSK) computer programme.

For overall cooling during laboratory experiments two "water fingers" were screwed to the lower side of the base plate. The water was temperature controlled at approx. 25^{o}C.

8. TEST SET-UP FOR 175-225MHz

In case of vision operation at peak sync output power levels beyond 400W a lot of typical measurements is required. More idealized, for a 3rd order transfer characteristic with phase distortion being neglected, a much faster and equipment saving method is the gain compression test. In that case there is an acceptable correlation to i.m.d., differential gain, cross-modulation and sync.compression.

Further, it can be said that it may be expected here that the conditions for typical television signals are somewhat relaxed because of the lower average power, decreasing again the junction temperatures within the BLV38.

9. TESTRESULTS OF SINGLE AND COMPLETE AMPLIFIER

To get an idea of the performance of the single amplifier branches it has been started without the coaxial hybrids.

Because the goal was to design broadband amplifiers they were once tuned for the most critical frequency 225MHz. So, no retuning since. Then they were combined and the complete unit measured again.

Both groups of results are summarized at pages 16 to 18.

10. CONCLUSIONS

The results of practical tests on a single BLV38 in a wideband circuit come up to the expections as are published in our Data Handbook (see 3.1.).

At the testfrequencies 175, 200 and 225MHz the output power is obtained and the compression about at the typical point. Gain and efficiency are typical.

Combining both amplifiers with 3dB-90^{o} hybrids results in an amplifier module with $\geq$440W-1dB compression and gain and efficiency about at the typical points too.

11. SURVEY OF TESTRESULTS

11.1. Single amplifier A:

f (MHz)	P_L (W)	P_F (W)	P_R (W)	P_S (W)	I_{C1} (W)	I_{C2} (A)	G_{p1} (A)	G_{p2} (dB)	eff. (%)	compr. (dB)
225	225	29.6	0.32	29.3	5.5	5.4	8.81	8.86	59.0	-0.66
225	200	25.5	0.33	25.2	5.1	4.9	8.94	9.00	57.1	-0.53
225	175	21.0	0.33	20.7	4.7	4.6	9.21	9.28	53.8	-0.26
225	150	18.0	0.30	17.7	4.3	4.2	9.21	9.28	50.4	-0.26
225	100	11.3	0.22	11.1	3.5	3.2	9.47	9.55	42.0	0
200	225	30.0	5.9	24.1	6.3	6.1	8.75	9.70	51.8	-0.46
200	200	26.4	5.2	21.2	5.8	5.6	8.80	9.85	49.9	-0.42
200	175	21.7	4.2	17.5	5.1	5.5	9.07	10.00	47.2	-0.14
200	150	18.8	3.7	15.1	5.0	4.7	9.02	9.97	44.4	-0.19
200	100	12.0	2.4	9.6	4.0	3.8	9.21	10.18	36.6	0
175	225	24.2	3.75	20.5	6.1	5.4	9.68	10.41	55.9	-0.88
175	200	19.5	3.2	16.3	5.6	4.9	10.11	10.89	54.9	-0.45
175	175	17.2	2.75	14.5	5.1	5.5	10.08	10.83	47.2	-0.48
175	150	14.0	2.25	11.7	4.7	4.2	10.30	11.06	48.4	-0.26
175	100	8.8	1.45	7.4	3.8	3.4	10.56	11.34	39.7	0

Note:

$P_S=P_F-P_R$

$G_{P1}=10 \log P_L/P_F$ dB

$G_{P2}=10 \log P_L/P_S$ dB

$V_{CC}=35V$

$I_{C(zs)}=2x200mA$

11.2. Single amplifier B:

f (MHz)	P_L (W)	P_F (W)	P_R (W)	P_S (W)	I_{C1} (A)	I_{C2} (A)	G_{p1} (dB)	G_{p2} (dB)	eff. (%)	compr. (dB)
225	225	31	1.8	29.2	5.1	5.1	8.61	8.87	63.0	-0.60
225	200	26	1.6	24.4	4.7	4.7	8.86	9.14	60.8	-0.35
225	175	22	1.4	20.6	4.35	4.35	9.01	9.29	57.5	-0.20
225	150	18.6	1.2	17.4	4.0	4.0	9.07	9.36	53.6	-0.14
225	100	12	0.75	11.25	3.3	3.25	9.21	9.49	43.6	0
200	225	30	3.8	26.2	5.8	6.1	8.75	9.34	54.0	-0.70
200	200	24.6	3.3	21.3	5.35	5.55	9.10	9.73	52.4	-0.31
200	175	21	2.9	18.1	5.0	5.1	9.21	9.85	49.5	-0.19
200	150	18	2.4	15.6	4.7	4.7	9.21	9.83	45.6	-0.21
200	100	11.5	1.6	9.9	3.75	3.8	9.39	10.04	37.8	0
175	225	26	5.5	20.5	5.35	5.9	9.37	10.40	57.1	-0.63
175	200	22	4.7	17.3	4.95	5.45	9.59	10.63	54.9	-0.41
175	175	18.8	4.0	14.8	4.6	5.0	9.69	10.73	52.1	-0.31
175	150	15.5	3.3	12.2	4.25	4.5	9.86	10.90	49.0	-0.14
175	100	10	2.25	7.75	3.5	3.7	10.0	11.11	39.7	0

Note:

$P_S = P_F - P_R$

$G_{p1} = 10 \log P_L/P_F$ dB

$G_{p2} = 1- \log P_L/P_S$ dB

$V_{CC} = 35V$

$I_{C(zs)} = 2x200mA$

11.3. Complete amplifier

f (MHz)	P_L (W)	P_F (W)	P_R (W)	P_S (W)	A I_{C1+C2} (A)	B I_{C1+C2} (A)	G_{p1} (dB)	G_{p2} (dB)	eff. (%)	compr. (dB)
225	100	11.3	0.16	11.14	4.95	4.20	9.47	9.53	31.2	0
225	200	22.6	0.34	22.30	6.95	6.00	9.47	9.54	44.1	0
225	250	28.3	0.43	27.90	7.60	6.75	9.46	9.53	49.8	-0.01
225	300	35.0	0.53	34.50	8.50	7.5	9.33	9.40	53.6	-0.14
225	350	42.0	0.71	41.30	9.05	8.15	9.21	9.28	58.1	-0.25
225	400	52.4	0.83	51.60	9.85	9.15	8.83	8.90	60.2	-0.64
225	450	68.0	1.20	66.80	10.75	10.20	8.21	8.28	61.4	-1.26
225	475	85.3	1.36	83.90	12.00	11.45	7.46	7.53	57.9	-2.01
200	100	12.6	0.1	12.50	5.95	5.80	9.00	9.03	24.3	0
200	200	24.9	0.95	24.00	8.35	8.20	9.05	9.22	34.5	+0.05
200	250	31.3	0.24	31.00	9.50	9.20	9.02	9.06	38.2	+0.02
200	300	37.6	0.29	37.30	10.30	10.10	9.02	9.05	42.0	+0.02
200	350	46.2	0.36	45.90	11.30	11.05	8.79	8.83	44.7	-0.21
200	400	54.3	0.41	53.90	12.30	12.1	8.67	8.71	46.8	-0.33
200	450	66.5	0.47	66.00	13.40	13.2	8.30	8.33	48.4	-0.70
175	100	9.34	0.12	9.22	5.60	4.90	10.30	10.35	27.2	0
175	200	18.50	0.24	18.26	7.90	6.85	10.34	10.40	38.7	+0.04
175	250	23.90	0.30	23.60	8.90	7.75	10.20	10.25	42.9	-0.10
175	300	28.6	0.37	28.23	9.80	8.50	10.21	10.26	46.8	-0.09
175	350	35.1	0.43	34.67	10.65	9.25	9.99	10.04	50.3	-0.31
175	400	43.1	0.56	42.54	11.70	10.40	9.68	9.73	51.6	-0.62
175	450	51.7	0.65	51.05	12.55	11.15	9.40	9.45	54.2	-0.90
175	475	63.3	0.85	62.45	13.75	12.35	8.75	8.81	52.0	-1.55

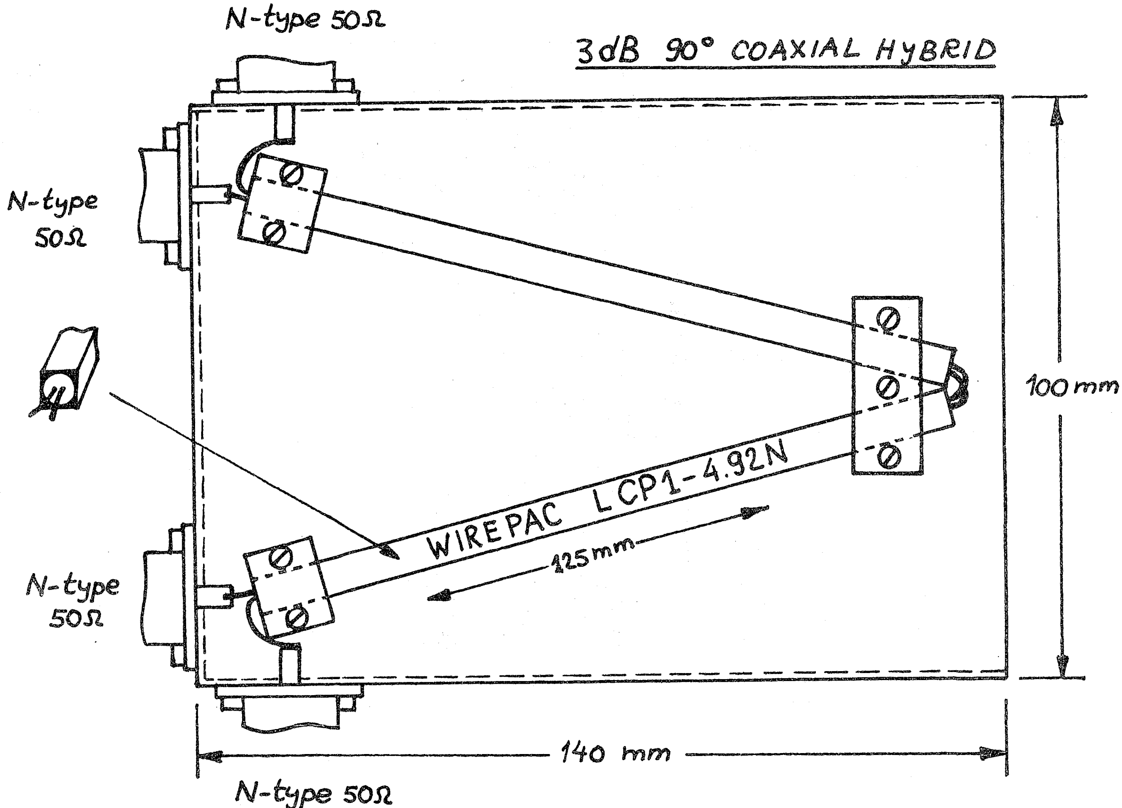

N-type 50Ω
3dB 90° COAXIAL HYBRID
N-type 50Ω
100 mm
WIREPAC LCP1-4.92N
125mm
N-type 50Ω
140 mm
N-type 50Ω

Fig.15

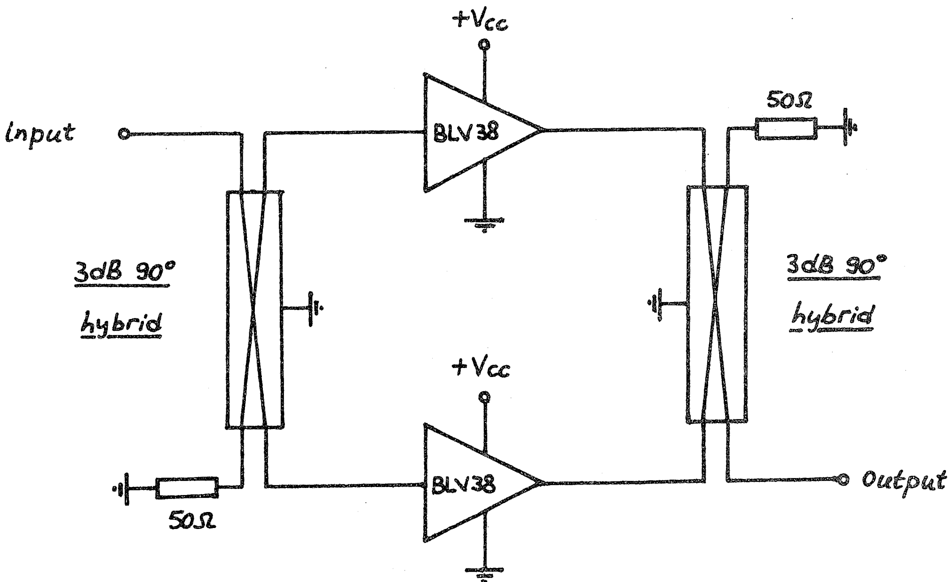

+Vcc
Input
BLV38
50Ω
3dB 90° hybrid
3dB 90° hybrid
+Vcc
BLV38
Output
50Ω

Fig.16

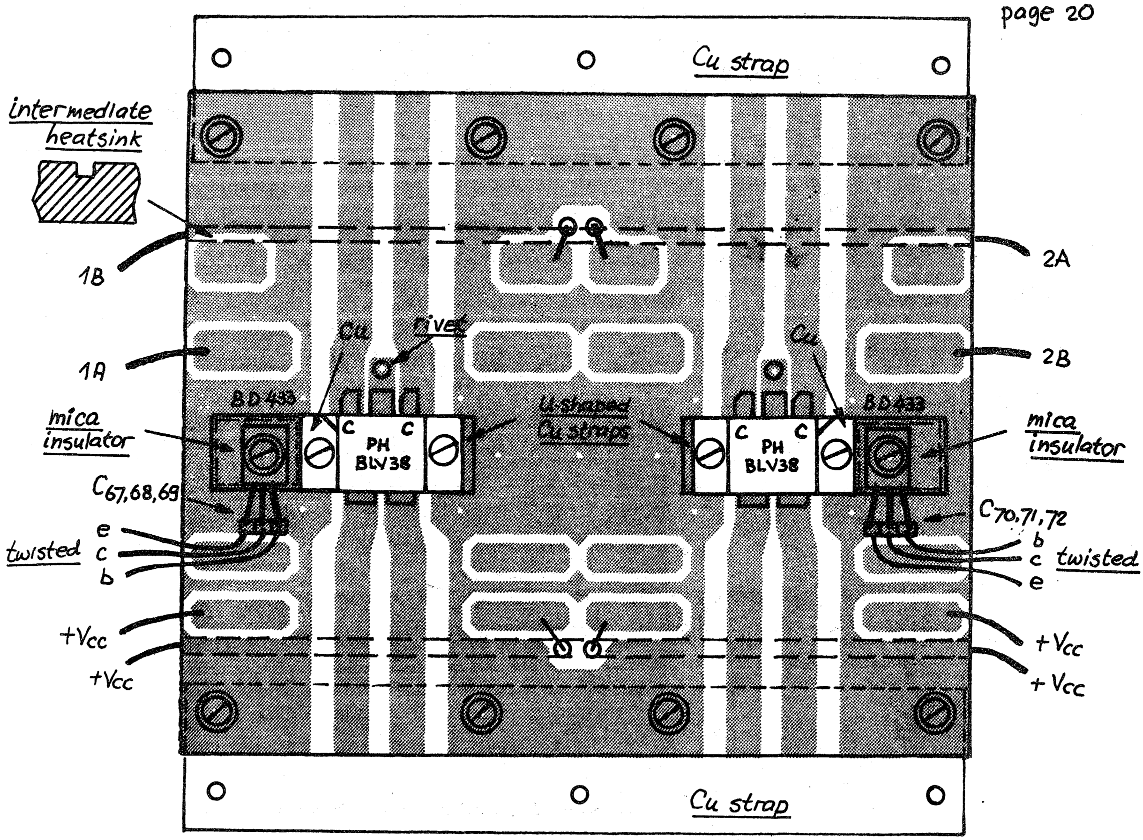

Cu strap
intermediate heatsink
1B
1A
2A
2B
Cu
rivet
Cu
BD 433
BD 433
mica insulator
mica insulator
U-shaped Cu straps
C
C
PH
BLV38
C
C
PH
BLV38
$C_{67,68,69}$
$C_{70,71,72}$
e
c
b
twisted
b
c
e
twisted
+Vcc
+Vcc
+Vcc
+Vcc
Cu strap

Fig.17

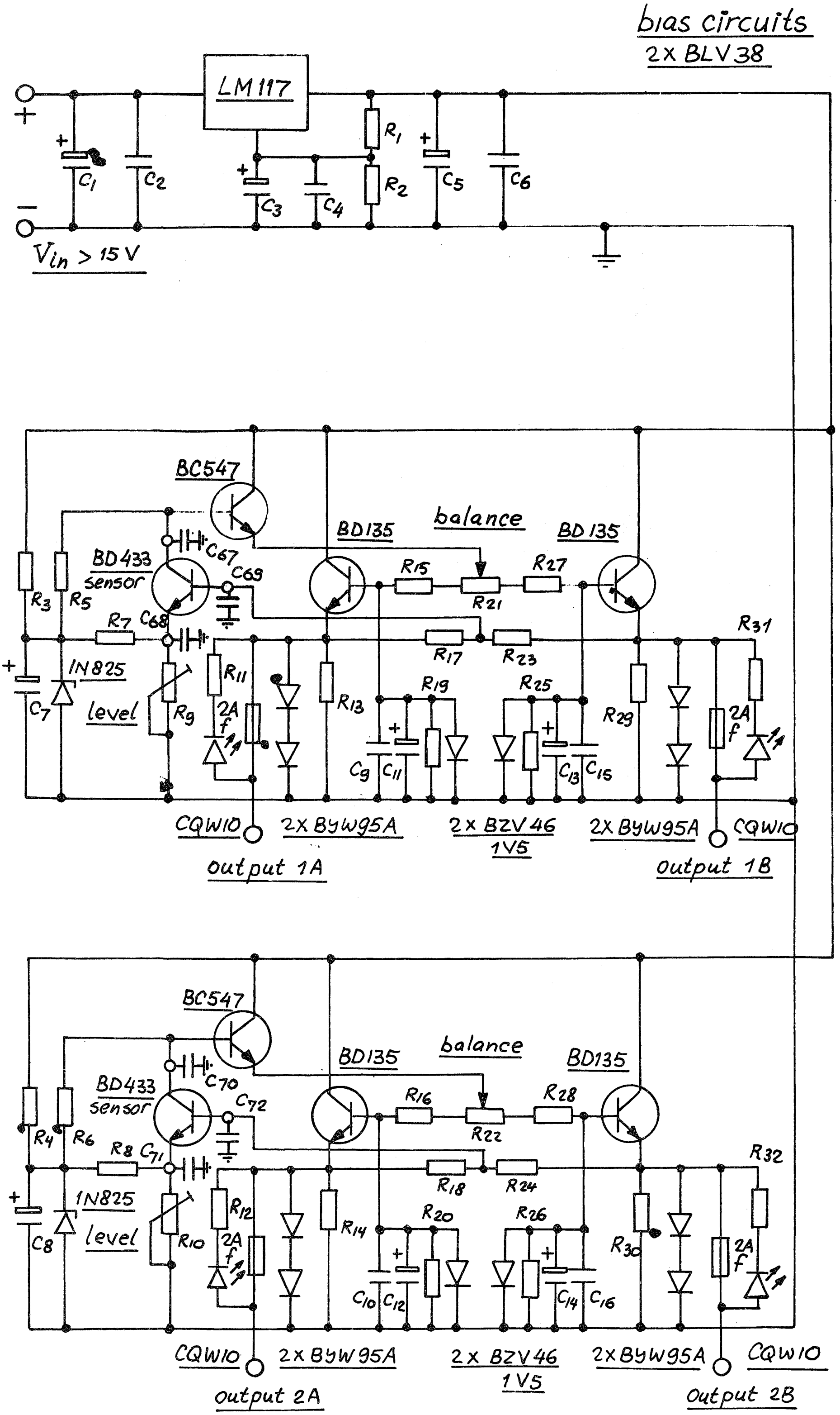
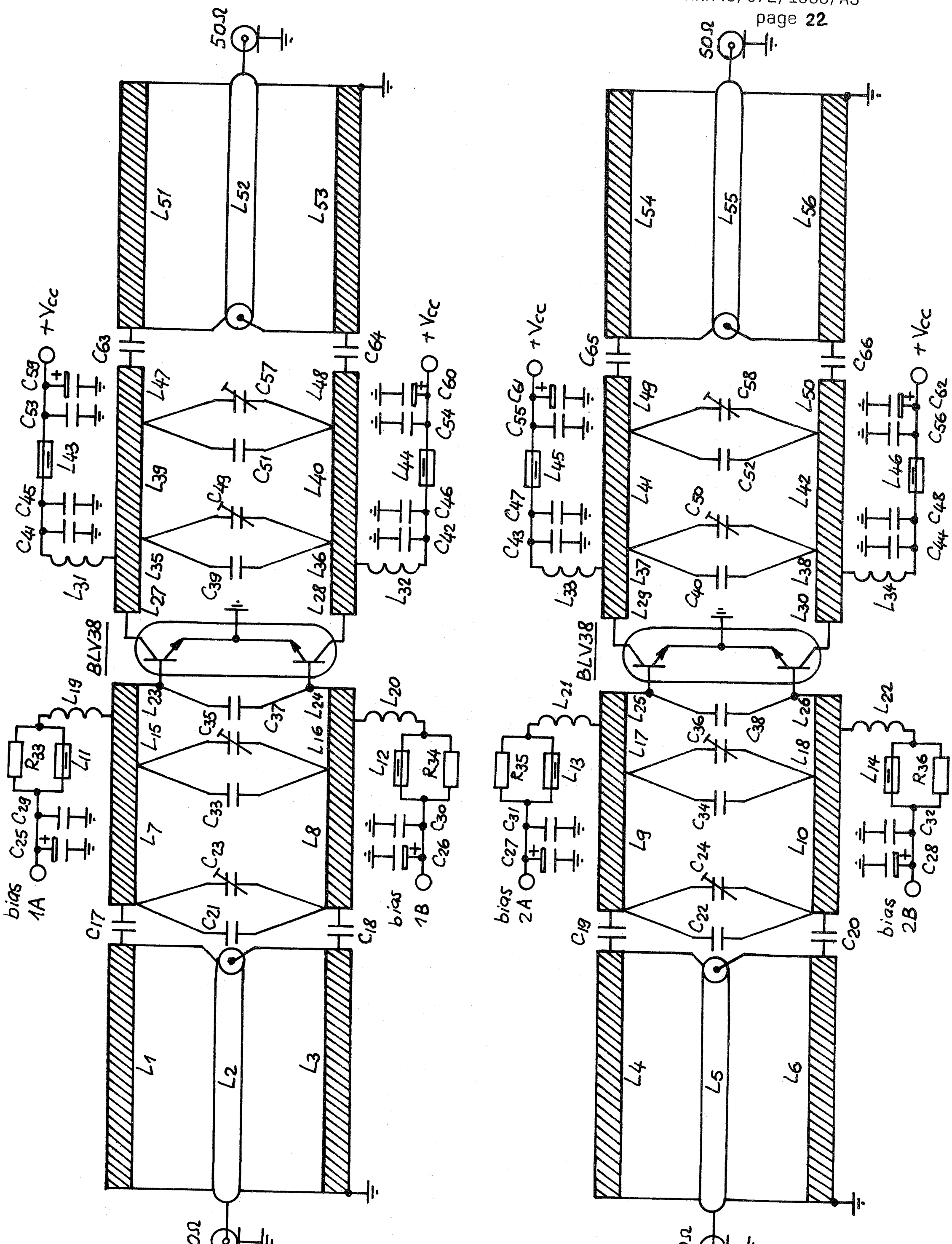

bias circuits
2 x BLV 38
LM117
V_{in} > 15 V
R_1
R_2
C_1
C_2
C_3
C_4
C_5
C_6
BC547
BD135
balance
BD 135
BD 433 sensor
C_{67}
C_{69}
C_{68}
R_3
R_5
R_7
R_{15}
R_{21}
R_{27}
R_{17}
R_{23}
R_{31}
1N825
level
R_9
R_{11}
2A f
R_{13}
R_{19}
R_{25}
R_{29}
C_7
C_9
C_{11}
C_{13}
C_{15}
CQW10
2 x BYW95A
2 x BZV46 1V5
2 x BYW95A
CQW10
output 1A
output 1B
BC547
BD135
balance
BD135
BD433 sensor
C_{70}
C_{72}
R_4
R_6
R_8
C_{71}
R_{16}
R_{22}
R_{28}
R_{18}
R_{24}
R_{32}
1N825
level
R_{10}
R_{12}
2A f
R_{14}
R_{20}
R_{26}
R_{30}
C_8
C_{10}
C_{12}
C_{14}
C_{16}
CQW10
2 x BYW 95A
2 x BZV46 1V5
2 x BYW95A
CQW10
output 2A
output 2B

Fig. 18


50Ω
50Ω
50Ω
50Ω
L1
L2
L3
L4
L5
L6
L7
L8
L9
L10
L11
L12
L13
L14
L15
L16
L17
L18
L19
L20
L21
L22
L23
L24
L25
L26
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L41
L42
L43
L44
L45
L46
L47
L48
L49
L50
L51
L52
L53
L54
L55
L56
C17
C18
C19
C20
C21
C22
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C24
C25
C26
C27
C28
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C31
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C43
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C44
C45
C46
C47
C48
C49
C50
C51
C52
C53
C54
C55
C56
C57
C58
C59
C60
C61
C62
C63
C64
C65
C66
R33
R34
R35
R36
BLV38
BLV38
bias 1A
bias 1B
bias 2A
bias 2B
+Vcc
+Vcc
+Vcc
+Vcc

Fig. 19

12. **LIST OF COMPONENTS**

C1=C3=C5=C7=C8=C11=C12=C13=C14=2.2μF(35V) tantalum electrolytic capacitor
C2=C4=C6=C9=C10=C15=C16=10nF±10%(500V)
C17=C18=C19=C20=2x27pF (500V) multilayer ceramic chip capacitors in parallel *
C21=C22=33pF (500V) multilayer ceramic chip capacitor *
C23=C24=2 to 18pF film dielectric trimmer (cat.no.2222 809 05003)
C25=C26=C27=C28=4.7μF (63V) electrolytic capacitor
C29=C30=C31=C32=C53=C54=C55=C56=10nF (50V) multilayer ceramic chip capacitor
C33=C34=47pF (500V) multilayer ceramic chip capacitor *
C35=C36=5 to 60pF film dielectric trimmer (cat.no.2222 809 08003)
C37=C38=2x100pF (500V) multilayer ceramic chip capacitors in parallel *
C39=C40=4x20pF (500V) multilayer ceramic chip capacitors in parallel *
C41=C42=C43=C44=C45=C46=C47=C48=470pF (500V) multilayer ceramic chip capacitor *
C49=C50=C57=C58=4 to 40pF film dielectric trimmer (cat.no. 2222 809 08002)
C51=C52=3x8.2pF (500V) multilayer ceramic chip capacitors in parallel *
C59=C60=C61=C62=10μF (63V) electrolytic capacitor
C63=C64=C65=C66=3x18pF (500V) multilayer ceramic chip capacitors in parallel *
C67=C68=C69=C70=C71=C72=330pF (50V) multilayer ceramic chip capacitor

L1=L3=L4=L6=L51=L53=L54=L56=50Ω stripline (4.8mmx80mm)
L2=L5=L52=L55=50Ω semi-rigid cable; outer dia.3.6mm; outer conductor length 80mm; soldered on striplines L3,L6,L53,L56.
L7=L8=L9=L10=43Ω stripline (6.0mmx28mm)
L11=L12=L13=L14=Ferroxcube wide-band HF choke; grade 3B (cat.no. 4312 020 36642)
L15=L16=L17=L18=43Ω stripline (6.0mmx6.0mm)
L19=L20=L21=7 turns closely wound enamelled Cu wire (0.4mm); int.dia.3mm; leads 2x4mm
L23=L24=L25=L26=L27=L28=L29=L30=L47=L48=L49=L50=43Ω stripline (6.0mmx12mm)
L31=L32=L33=L34=30nH; 2 turns enamelled Cu wire (1.6mm); int.dia.6mm; leads 2x3mm; coillength 5mm; connected 12mm from transistor edge.
L35=L36=L37=L38=43Ω stripline (6.0mmx4mm)
L39=L40=L41=L42=43Ω stripline (6.0mmx29mm)
L43=L44=L45=L46=Ferroxcube wide-band HF choke; grade 3B (cat.no.4312 020 36642), 2 straight wires (0.8mm) through FXC-bead in parallel.

R1=2KΩ
R2=240Ω
R3=R4=560Ω
R5=R6=39KΩ
R7=R8=4.7KΩ
R9=R10=500Ω variabel 10 turns
R11=R12=R31=R32=1K8-0.5W (depends on V_{CC}; in this case V_{CC}=35V)
R13=R14=R29=R30=68Ω
R15=R16=R27=R28=180Ω
R17=R18=R23=R24=680Ω
R19=R20=R25=R26=1KΩ
R21=R22=100Ω variabel 10 turns

The several striplines are on a double Cu-clad printed circuit board with glass microfibre reinforced PTFE, dielectric (ϵ_r=2.2); thickness 1/16"; thickness of copper sheet 2x35μm.

The circuit and the components are on one side of the p.c. board, the other side is unetched copper to serve as ground plane.
Earth connections are made by copper straps under the emitter lead and at several places as has been shown in Fig.19.
Also hollow rivets and a lot of M2 and M3 screws are applied.

* American Technical Ceramics (ATC) capacitor type 100B or capacitor of same quality.

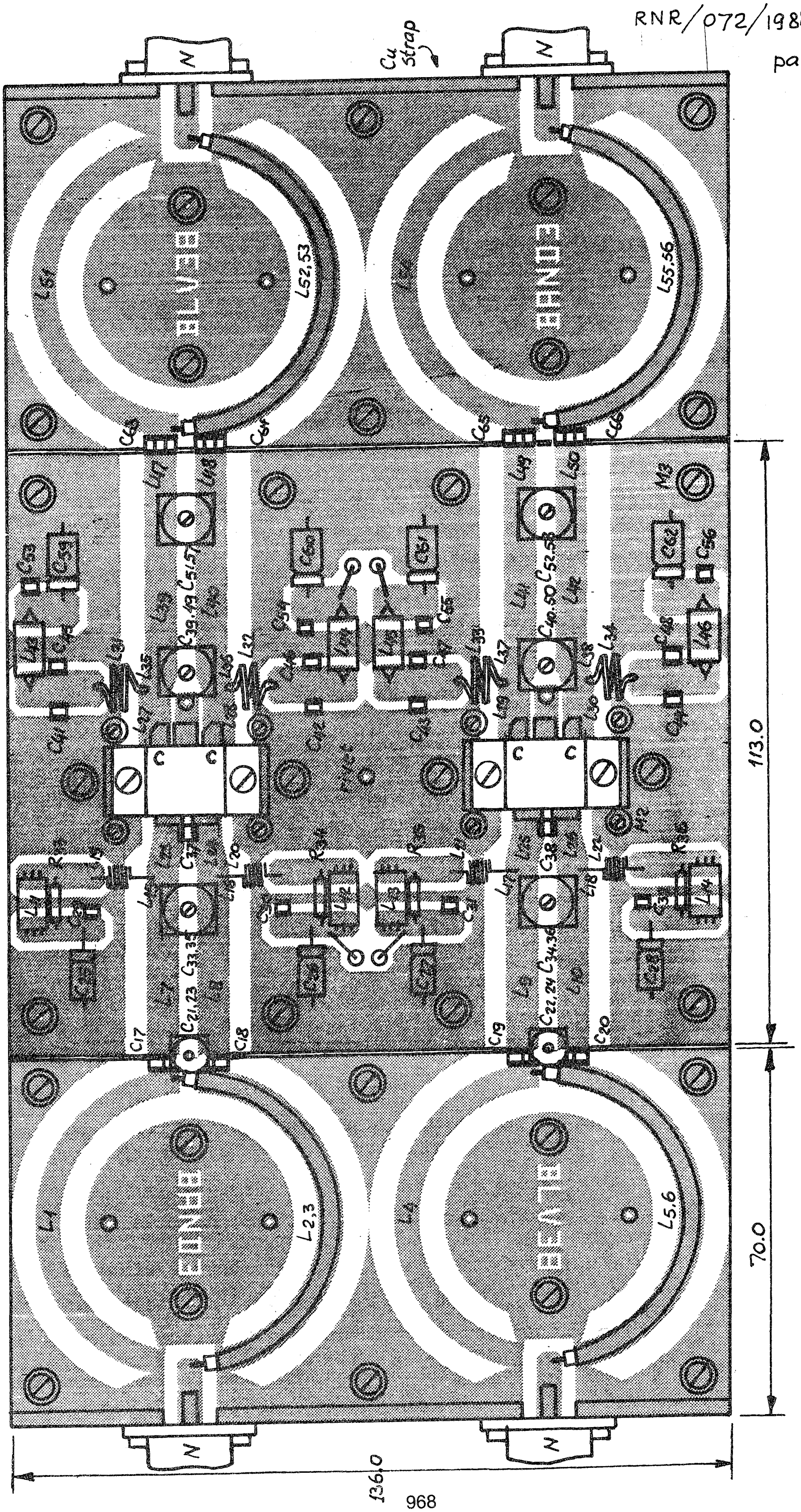

Fig.20

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