

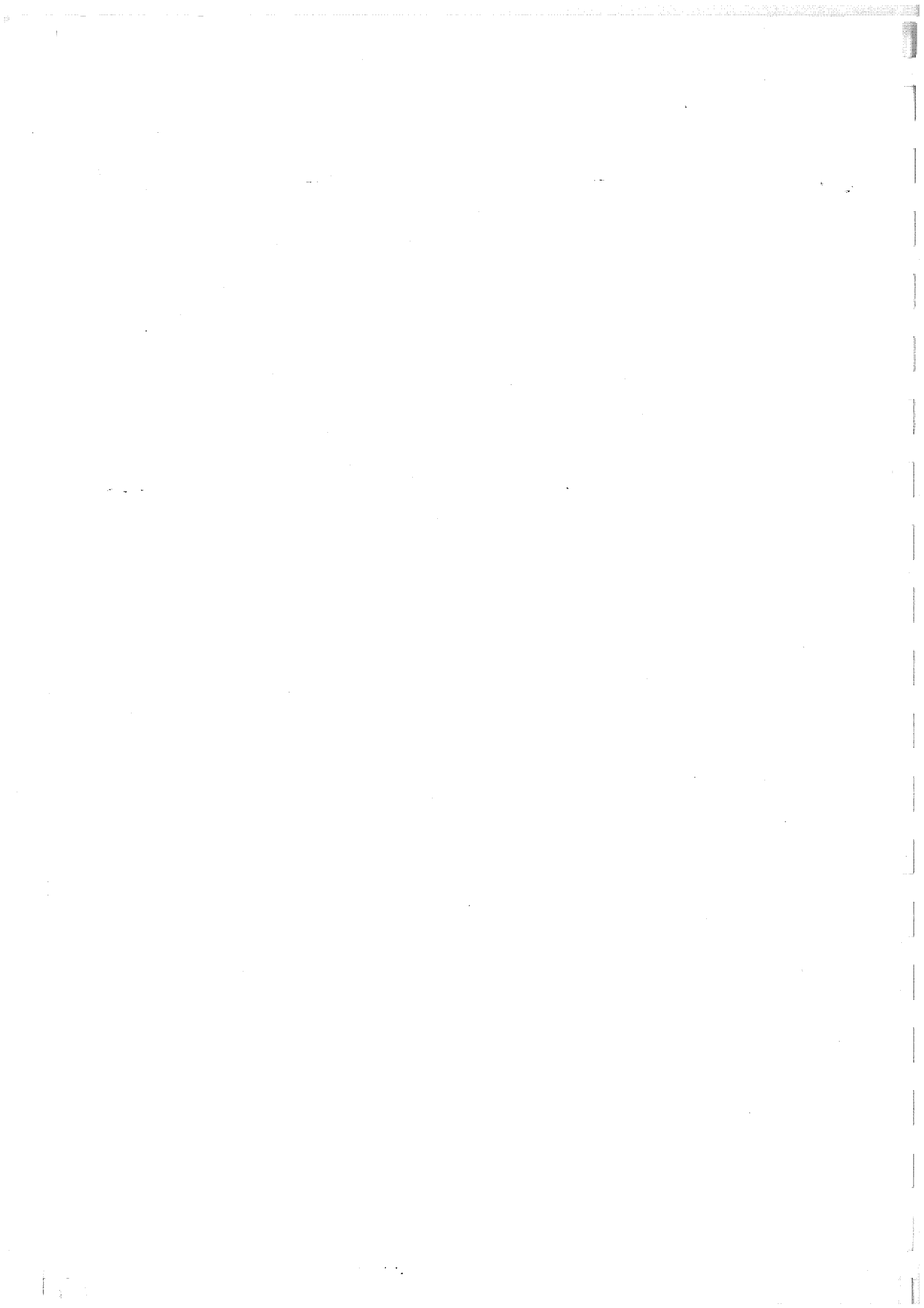


LEVEL MEASURING SET PSM-5  
Frequency range : 10 kHz to 36 MHz  
Service Manual BN 444

1.8.68 Bi  
0.15.3.74 2109 UN v. 1594

Subject to change without notice

**Wandel u. Goltermann - Reutlingen/Germany**



# CONTENTS

1. INTRODUCTION . . . . .	1
2. DETAILED INDIVIDUAL SUBUNIT DESCRIPTION . . . . .	3
2.1. Control Oscillator RWO-5 . . . . .	3
2.1.1. Control Oscillator (2) and Control Frequency Buffer (1) . . . . .	3
2.1.2. Function of coil L 103 at the Control Frequency Output for the Receiver . . . . .	4
2.1.3. Frequency Converter (3) . . . . .	5
2.1.4. Coarse Frequency Oscillator (4) . . . . .	6
2.1.5. Auxiliary Oscillator (5) . . . . .	7
2.1.6. Discriminator for Auxiliary Oscillator (6) . . . . .	8
2.1.7. Linearizing Network (7) . . . . .	8
2.1.8. Frequency and Phase Control Circuit . . . . .	10
2.1.9. Counting Discriminator I (9) . . . . .	11
2.1.10. Counting Discriminator II (10) . . . . .	11
2.1.11. Standard Frequency Buffer (11) . . . . .	12
2.1.12. 10 : 1 Frequency Divider (12) . . . . .	12
2.1.13. Standard Frequency Oscillator (13) . . . . .	14
2.1.14. Phase Meter (14) . . . . .	15
2.1.15. Lock-In Indicating Amplifier (15) . . . . .	16
2.1.16. Saw-Tooth Generator (16) . . . . .	18
2.1.17. 2.7 MHz Crystal Oscillator (17) . . . . .	20
2.1.18. Interpolation Oscillator (18) . . . . .	21
2.1.19. Tuning Indication Circuit (19) . . . . .	22
2.1.20. Frequency Converter (20) , Buffer and Switching Stage (21) . . . . .	23
2.1.21. 10 : 1 Frequency Divider (22) . . . . .	25
2.1.22. Discriminator (23) . . . . .	25
2.1.23. Power Supply (24) . . . . .	26
2.2. Generator Portion PS-5 . . . . .	32
2.2.1. 69 MHz Oscillator (1) . . . . .	33
2.2.2. Phase Control (2) . . . . .	33

2.2.3. Amplitude Modulator (3)	37
2.2.4. 69 MHz Amplifier (4)	38
2.2.5. Frequency Converter (5)	39
2.2.6. 36 MHz Low-pass Filter (6)	39
2.2.7. Pre-amplifier (7)	39
2.2.8. Output Amplifier (8)	40
2.2.9. Attenuator (9)	42
2.2.10. Amplitude Control (10)	42
2.2.11. Limiter (11)	44
2.2.12. Voltage Filtering (12)	44
2.2.13. Hunting Oscillator (13)	44
2.3. Receiver Portion PM-5	46
2.3.1. Input Circuit (1)	47
2.3.2. Calibration Oscillator 100 kHz (2)	49
2.3.3. Pre-amplifier 10 kHz to 36 MHz (3)	49
2.3.4. 36 MHz Low-pass Filter (4)	50
2.3.5. Mixer I and 69 MHz Amplifier (5)	50
2.3.6. Limiter 69.01 to 105 MHz (6)	51
2.3.7. 69 MHz Band-pass Filter (7)	52
2.3.8. Mixer II and Buffer Stage (8)	52
2.3.9. 7.44 MHz Band-pass Filter (10)	53
2.3.10. Mixer III (11)	53
2.3.11. 600 kHz Band-pass Filter (11) (12)	53
2.3.12. Mixer IV (12)	54
2.3.13. 30 kHz Band-pass Filter (14)	54
2.3.14. 30 kHz Amplifier I and Post-attenuator (15)	54
2.3.15. 30 kHz Amplifiers 2 and 3 (16)	55
2.3.16. Output Network (16)	56
2.3.17. Logarithmic circuit (16)	57
2.3.18. Converter Oscillator for Single Sideband Demodulation (17)	58
2.3.19. Switch Assembly and Level Switch (18)	59
2.3.20. Main Distributor (19)	59



2.3.21. Distributor 2 (20)	59
2.3.22. Generating of Carrier Frequencies II to IV	60
2.3.22.1. Frequency Locking Device for 10/570 kHz (13)	60
2.3.22.2. 570 kHz Crystal Oscillator (12)	61
2.3.22.3. Frequency Locking Device for 1.71/6.84 MHz (12)	62
2.3.22.4. 6.84 MHz Oscillator (11)	63
2.3.22.5. Frequency locking device for 20.52/61.56 MHz (10)	63
2.3.22.6. Oscillator and Limiter for 61.56 MHz (9)	64
2.4. Scale Expander AZD-1	65
2.5. Probe TK-8	67
2.6. Return Loss Accessory RFZ-5	68
2.6.1. Measuring principle	68
2.6.2. Circuit	69
2.6.3. Reflection Standards	70
2.6.4. Measuring Cable	71
3. LIST OF THE REQUIRED MEASURING INSTRUMENTS FOR FAULT FINDING AND TESTING OF THE UNITS	73
4. FAULT FINDING	75
4.1. Oscillator Portion RWO-5	75
4.1.1. Cabinet Removal and Access to the Sub-Units	75
4.1.2. Faults in the Power Supply	77
4.1.3. Malfunction of the Coarse Frequency Oscillator Locking Indicator	77
4.1.4. Malfunction of the Interpolation Oscillator Tuning Indicator	79
4.1.5. Malfunction in the Generation of the Control Frequency	80
4.2. Generator PS-5	83
4.2.1. Complete Failure of the Unit	83
4.2.1.1. Checking the Operating Voltages	83
4.2.1.2. Checking of the Carrier Frequency Supply	83
4.2.2. Checking of Limiter (11)	84
4.2.3. Checking of the Frequency Converter (5)	87
4.2.4. Malfunction in the Combination 69 MHz Oscillator, Amplitude Modulator and 69 MHz Amplifier	87

4.2.4.1. Checking the 69 MHz Oscillator . . . . .	87
4.2.4.2. Checking the 69 MHz amplifier . . . . .	87
4.2.4.3. Instructions for Fault Finding in the 69 MHz Amplifier . . . . .	87
4.2.5. Malfunction in the Combination of Pre-amplifier and Final amplifier . . . . .	89
4.2.5.1. Amplification of the Pre-amplifier . . . . .	89
4.2.5.2. Amplification of the Final Amplifier . . . . .	89
4.2.6. Failure of the Phase Control . . . . .	90
4.2.6.1. Checking the Synchronizing Frequency . . . . .	90
4.2.6.2. Checking the Hunting Oscillator . . . . .	90
4.2.6.3. Checking the 69 MHz Signal . . . . .	90
4.2.6.4. Checking the Phase Control . . . . .	90
4.2.6.5. Hold and Pull in Range of the Phase Control . . . . .	92
4.2.7. Malfunction of the Indicator Circuit or Control Amplifier . . . . .	92
4.2.7.1. Operational Check of the Indicator circuit and the Control Amplifier . . . . .	92
4.2.8. Malfunction of the Amplitude Control . . . . .	93
4.2.8.1. Operational Check of the Amplitude Control . . . . .	93
4.2.8.2. Control Amplification of the Amplitude Control . . . . .	94
4.2.9. Malfunction of the Attenuator . . . . .	94
4.3. Receiver PM-5 . . . . .	98
4.3.1. Operating Voltages . . . . .	100
4.3.2. Test Points for the Rapid Localization of Faults . . . . .	100
4.3.2.1. Examination of the Levels . . . . .	100
4.3.2.2. Checking the Frequency Locking . . . . .	101
4.3.3. Gas Pressure Relays . . . . .	104
4.3.3.1. Sensitivity Selection . . . . .	105
4.3.3.2. Linear and Logarithmic Calibration . . . . .	105
4.3.3.3. Bandwidth . . . . .	105
4.3.3.4. Demodulation Modes . . . . .	105
4.3.3.5. Power Level . . . . .	106
4.3.4. Mixer I Carrier Balance . . . . .	106
4.3.5. Frequency Response . . . . .	107
4.3.5.1. $Z = 75 \Omega$ . . . . .	107
4.3.5.2. $Z = 50, 60, 65, 135$ and $150 \Omega$ . . . . .	109
4.3.5.3. High Impedance Input " $R_e = 1/\omega C_e$ " ( $Z_{in} = 1/\omega C_{in}$ ) . . . . .	110
4.3.6. Attenuation Values . . . . .	110
4.3.6.1. Attenuation of Distortion . . . . .	110
4.3.6.2. IF Suppression . . . . .	111
4.3.6.3. Image Rejection . . . . .	111

4.3.7. Disturbing Signals . . . . .	111
4.3.7.1. Intrinsic Noise . . . . .	111
4.3.7.2. Interfering Voltages in the Absence of an Input Signal . . . . .	112
4.3.7.3. Interfering Voltages on Sweeping with the Input Signal . . . . .	114
4.3.8. Transfer Impedance . . . . .	115
5. CHECKING AND ALIGNMENT . . . . .	123
5.1. Oscillator Portion RWO-5 . . . . .	123
5.1.1. Instructions Regarding the Checking of the Power Supply . . . . .	123
5.1.2. Power Supply . . . . .	123
5.1.3. Checking of the two point regulator . . . . .	125
5.1.4. Residual AC measurement . . . . .	126
5.1.5. Checking the 1 MHz Standard Frequency . . . . .	126
5.1.6. Checking the Output Signals of the Standard Frequency Buffer . . . . .	127
5.1.7. Checking the Divided Frequencies . . . . .	127
5.1.8. Checking the Frequency of the 2.7 MHz Oscillator . . . . .	127
5.1.9. Checking the Frequency of the Interpolation Oscillator, the 10 : 1 Frequency Divider and that of the Frequency Converter . . . . .	127
5.1.10. Checking the Carrier Level for the Frequency Converter . . . . .	128
5.1.11. Operational Check of the Phasemeter . . . . .	128
5.1.12. Checking the Output Amplitude of the Auxiliary Oscillator . . . . .	129
5.1.13. Adjustment and Checking of the Control Frequency Amplitude . . . . .	129
5.1.14. Checking the Signal Level for Frequency Converter . . . . .	130
5.1.15. Adjustment of the Discriminator Current and Alignment of the Linearization . . . . .	130
5.1.16. Characteristic Curve Alignment of Counter Discriminator I . . . . .	131
5.1.17. Checking the Frequency and Phase Control . . . . .	131
5.1.18. Checking of the Hold Range of the Phase Control . . . . .	132
5.1.19. Checking the Output Amplitude of the Sawtooth Generator . . . . .	132
5.2. Generator Portion PS-5 . . . . .	136
5.2.1. Operating Voltages . . . . .	136

5.2.2.	Alignment of the 69 MHz Oscillator and Phase Control . . . . .	136
5.2.3.	Alignment of the 69 MHz Amplifier . . . . .	137
5.2.4.	Adjustment of P 1003 . . . . .	138
5.2.5.	Checking the Adjustment Range of P 1002 . . . . .	138
5.3.	Receiver Portion PM-5 . . . . .	138
5.3.1.	Terminating Resistors . . . . .	138
5.3.2.	10 dB (4 Np) Pre-Attenuator and Input Capacitance . . . . .	140
5.3.2.1.	Alignment of the Attenuator . . . . .	140
5.3.2.2.	Input Capacitance . . . . .	142
5.3.3.	Cathode Follower Stage . . . . .	142
5.3.4.	Calibration Oscillator . . . . .	142
5.3.5.	Pre-amplifier . . . . .	142
5.3.6.	36 MHz Low-pass Filter . . . . .	144
5.3.7.	Limiter for 69.01 to 105 MHz . . . . .	144
5.3.8.	69 MHz Bandpass Filter . . . . .	144
5.3.9.	7.44 MHz Bandpass Filter, 600 kHz Bandpass Filter . . . . .	146
5.3.10.	61.56 MHz Oscillator and Phase Locking Device . . . . .	147
5.3.11.	6.84 MHz Oscillator and Phase Locking Device . . . . .	147
5.3.12.	30 kHz Bandpass Filter . . . . .	147
5.3.12.1.	Checking the selectivity curve of the PM-5, which essentially corresponds to that of the bandpass filter. . . . .	147
5.3.12.2.	Alignment of the 30 kHz bandpass filters . . . . .	150
5.3.12.3.	Insertion loss compensation . . . . .	150
5.3.13.	Logarithmic Circuitry . . . . .	150
5.3.13.1.	Alignment of C 1602 . . . . .	
5.3.13.2.	Logarithmic characteristic curve . . . . .	151
5.3.14.	Output Sockets . . . . .	151
5.3.15.	Frequency Conversion Oscillator for the Demodulation of SSB Signals . . . . .	151
5.3.16.	Voltage-free Front Panel . . . . .	153
5.4.	Scale Expander AZD-1 . . . . .	153
5.4.1.	Opening the Cabinet . . . . .	153
5.4.2.	Removing the Unit from its Cabinet . . . . .	154
5.4.3.	Current Consumption . . . . .	154

5.4.4. Checking the Attenuator . . . . .	. 155
5.4.5. Checking the Expansion Adjustment (P 3) . . . . .	. 155
5.4.6. Alignment of the Attenuator . . . . .	. 157
5.4.7. Expansion Adjustment . . . . .	. 157
5.5. Probe TK-8 . . . . .	. 157
5.5.1. Current Consumption and Operating Points . . . . .	. 157
5.5.2. Attenuation Alignment . . . . .	. 158
5.5.3. Frequency Response . . . . .	. 158
6. OSCILLOSCOPE TRACES . . . . .	. 163
6.1. Control Oscillator RWO-5 . . . . .	. 163
6.2. Generator PS-5 . . . . .	. 168
6.3. Receiver PM-5 . . . . .	. 170
7. ADJUSTMENT OF THE OPTICAL PORTION AND THE DRIVE OF THE OSCILLATOR PORTION . . . . .	. 173
7.1. Condenser and Lens . . . . .	. 173
7.2. Mirror Adjustment . . . . .	. 175
7.3. Drive Plate . . . . .	. 176
8. INSTRUCTIONS REGARDING MAINTENANCE AND SERVICE . . . . .	. 177
8.1. Switch Care . . . . .	. 177
8.2. Cleaning of the Lenses . . . . .	. 177
8.3. Exchange of the Projection Lamp . . . . .	. 177



## List of Instrument Illustrations

Fig. 23	Return Loss Measuring Accessory . . . . .	70
Fig. 24	Control Oscillator with Chassis swung open . . . . .	76
Fig. 35	PS-5 Top View . . . . .	96
Fig. 36	PS-5 Bottom View . . . . .	97
Fig. 37	PM-5 Top View . . . . .	99
Fig. 47	PSM-5 Power Supply . . . . .	124
Fig. 49	RWO-5 Bottom View . . . . .	133
Fig. 50	RWO-5 Top View . . . . .	134
Fig. 51	RWO-5 Rear View . . . . .	135
Fig. 56	PM-5 Input Network . . . . .	141
Fig. 60	PM-5 Rear View . . . . .	145
Fig. 62	PM-5 Side View . . . . .	148
Fig. 63	PM-5 Swing-Out Chassis (internal view) . . . . .	149
Fig. 64	PM-5 Intermediate Chassis . . . . .	152
Fig. 73	Coarse Tuning and Auxiliary Oscillator with complete drive . . . . .	174





Due to the advanced design and the interaction between the three individual units of the complete measuring set, it is not easy to establish the location of a fault. The section "DESIGN AND THEORY OF OPERATION" of the Operating Manual and especially section 2 "DETAILED INDIVIDUAL SUBUNIT DESCRIPTION" of this Maintenance Manual, serve to acquaint the user with the individual circuits of the complete measuring set.

Section 4, "FAULT LOCATION" of this maintenance manual gives instructions for systematic fault tracing from circuit to circuit by means of voltage tests and oscilloscope displays. Section 5 of this manual, "CHECKING AND ALIGNMENT" is also an aid in the above task, since this section explains the main tests and alignment tasks required of the individual units.

Section 6 contains all the oscilloscope displays required for a functional check of each individual instrument. These illustrations are also referred to in the section "FAULT FINDING".

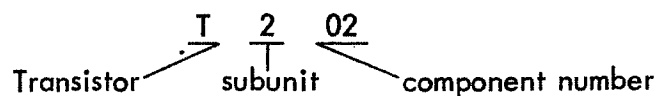
The section "MAINTENANCE AND REPAIR" contains instructions pertaining to the care of the complete measuring set.

The individual subunits referred to in this manual are numerically coded:

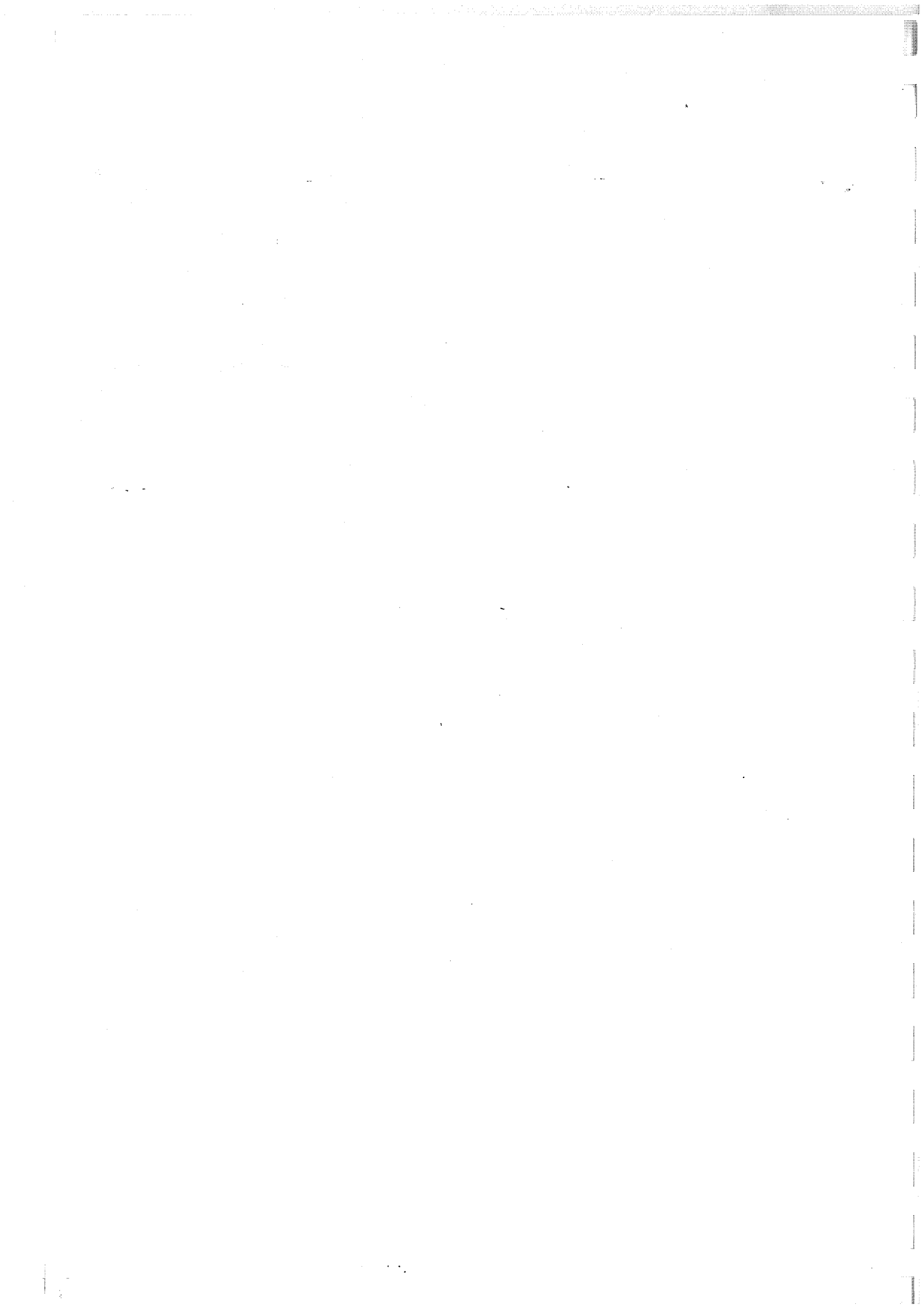
e.g. ② Control Oscillator.

Each component located in this unit has the number 2 as its initial digit. Each component is further defined by two following digits, such as 01 in resistor R 201 or inductance L 201.

Thus, if transistor 02 to be found in circuit diagram ② the coding will be as follows:



When ordering replacements, full information must include the order number of the measuring instrument, the serial number of the instrument and the full component number: e.g. BN 241 No. 36737 D T 202



## 2. DETAILED INDIVIDUAL SUBUNIT DESCRIPTION

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### 2.1. Control Oscillator RWO-5

#### 2.1.1. Control Oscillator (2) and Control Frequency Buffer (1)

The Generator PS-5 and the Receiver PM-5 derive the control signal required for synchronous tuning from the control oscillator via the buffer stage of the RWO-5. A frequency of 69 MHz serves as the fixed frequency for the generator and as the first intermediate frequency of the receiver. The frequency range of the control oscillator thus extends from 69.01 to 105 MHz corresponding to the signal frequency range of the measuring set from 10 kHz to 36 MHz. The control voltage for the generator and receiver is  $U_{\text{rms}} = 0.5 \text{ V}$  (load impedance = 75  $\Omega$ )

The full oscillator frequency determining capacitance consists of varicap diodes GI 201 to GI 208, whose capacitance is adjusted by means of the output voltage of the linearization circuit (7).

The resonant inductance consists of a wire loop accommodated in a metal housing and tuning is achieved by means of a metal tab. The oscillator output is fed to the pre-amplifier T 110/T 111 via the buffer stage T 202, which has no retro-action and which presents a resistive load to the oscillator. The subsequent output stage T 101/T 102 has an output impedance of 75  $\Omega$  and presents a non-load voltage of 1 V to output sockets Bu 103 and Bu 106.

The attenuated signal of the output stage is fed to frequency converter 3 at socket Bu 107 with an amplitude of 0.1 V at 75  $\Omega$  output impedance.

A DC voltage is derived by the rectifier circuit GI 101/GI 102 from the final stage output of T 101/T 102, and is compared with a reference voltage (voltage at P 101) in amplifier stage T 103. The output current of T 103 determines the emitter current of the transistor oscillator T 201 and thus the oscillator amplitude. This results in a near constant output signal throughout the whole frequency range.

Adjustment of potentiometer P 101 alters the reference voltage and as a result the oscillator amplitude.

The filter networks mounted on circuit boards 241-X and 241-Y of the control oscillate decouple the individual stages and filter the supplied operating voltage.

### 2.1.2. Function of coil L 103 at the Control Frequency Output for the Receiver

The coil L 103 is to be found between the sockets Bu 104 and Bu 105. It consists of a screened RF-cable wound on to a pot core.

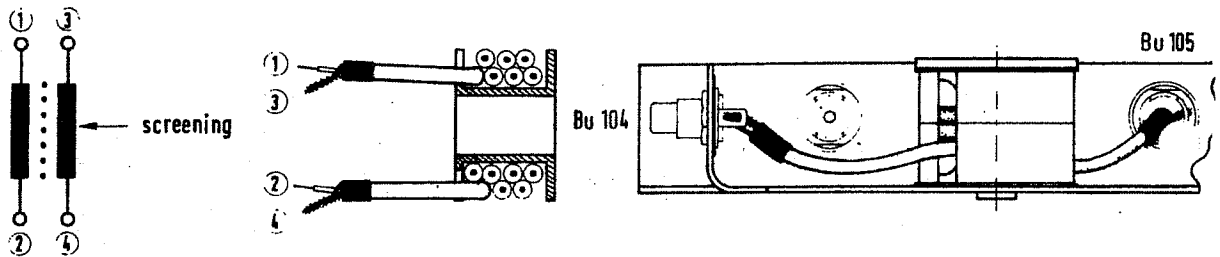


Fig. 1

The coil is designed to fulfill two tasks:

- 1) It must present a sufficiently high mutual impedance of  $\geq 7 \Omega$  between the equal-potential chassis of the RWO-5 and PS-5 to the Receiver portion PM-5, throughout the measuring frequency range of 10 kHz to 36 MHz. This is of importance when testing high loss objects since a ground shunt to the test object between generator and receiver would bring errors into the measurement.

The cable screening (turns 3 to 4) forms the desired impedance. The output current is thus forced to flow from the generator via the test object to the receiver and to return only via test object ground to the generator.

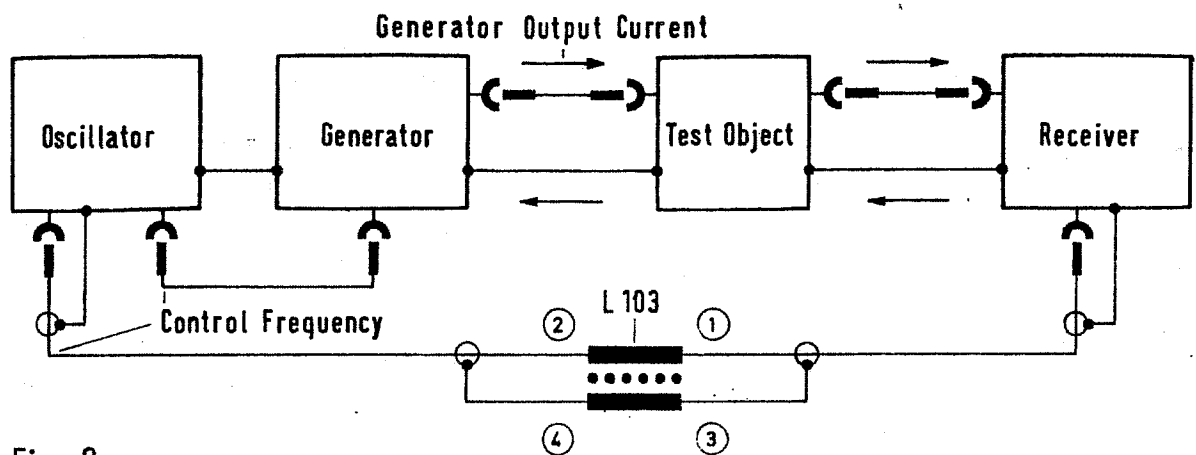


Fig. 2

- 2) It must at the same time present a low impedance control frequency connection between the Control Oscillator RWO-5 and the Receiver PM-5 so that the interference pick-up is avoided.

The internal and external conductors of the coiled RF-cable form a bifilar winding: i.e. the currents of the two conductors are arranged to oppose each other thus cancelling their effects.

This results in the same low-impedance for the control frequency as would be presented by a non-coiled cable.

### 2.1.3. Frequency Converter (3)

The task of this frequency converter is to form a difference frequency from the frequencies of the coarse frequency oscillator and the control oscillator. This difference frequency is required as a reference for the interpolation oscillator frequency. The coarse frequency oscillator signal then assumes the role of the "carrier" and is applied to the buffer stage T 301, located on circuit board 241-Z, at an amplitude of approx. 450 mV.

The input signal with an amplitude of approx. 90 mV is applied to the two buffer stages T 303 and T 304 mounted on circuit board 241-AS.

Since the frequency converter passes the difference frequency to a phase shifting network, it is important that a large amplitude difference be maintained between the residual carrier and the desired output signal to avoid disturbances during phase shifting. The low-pass filter network comprising L 302/C 305 and L 901/C 901 at the input of the counting discriminator 1 as well as the basic design of the converter meet the above requirement.

The 1 MHz or 100 kHz spikes produced in the phase meter (14) have a retro-action on the converter input (3) via buffer transistors T 1402 and T 1404. This, together with the coarse oscillator frequency or with the control frequency via the converter, may produce undesirable beat frequencies. Two additional buffer transistors T 305 and T 306 are provided to suppress these interference signals.

#### 2.1.4. Coarse Frequency Oscillator (4)

The coarse frequency oscillator can be phase-locked. Its frequency range covers 66 to 102 MHz. During tuning, the flywheel swings its frequency into the hold-in range of a sync. circuit which locks the frequency to a multiple of 100 kHz. The frequency setting is indicated on a projected scale having a calibration marker spacing of 100 kHz. Provision is made for locking the oscillator to any one of these markers. The oscillator consists of a three-point circuit with transistor T 401, inductance L 401 and variable capacitor C 407. Feed-back is achieved via a coil tapping and C 404 to the emitter of transistor T 401. The varicap GI 401 serves as the frequency control element.

To achieve this, the phase-meter (14) compares the oscillator signal with the phase position of a 100 kHz square wave voltage and presents the varicap with the voltage required for adjustment to the nominal value.

In the free-running oscillator mode, as selected by switch S 401, the output of the Zener diode GI 402 is applied to varicap GI 401 and is thus biased to 10 V. The frequency locking circuit is then disconnected and the oscillator can be tuned freely. The meter J 2302 receives an additional current corresponding to f.s.d. which in turn indicates the free-running mode of the oscillator.

The oscillator output is applied to frequency converter (3) and phase-meter (14) via two buffer transistors T 402 and T 403. The r.m.s. value is approx. 0.4 to 0.6 V.

Conditions regarding the Stability of the Phase-lock

Frequency deviations due to variations of temperature, shock or other effects must be kept to a minimum to ensure that the oscillator does not wander from the phase-locked condition. This requires a rugged design, a small temperature coefficient and a wide pull-in range. The circuit temperature coefficient is approx.  $-3 \times 10^{-5}/^{\circ}\text{C}$ , the pull-in range is approx.  $\pm 12$  kHz at 66 MHz and approx.  $\pm 35$  kHz at 102 MHz.

#### 2.1.5. Auxiliary Oscillator (5)

Due to the operating principle (see functional description of the oscillator RWO-5) the linearizing circuit (7) requires a control current which varies as a linear function of the coarse oscillator frequency. This should then cause the coarse frequency oscillator to be lead by the control oscillator by approx. 3 MHz.

The difficulties of designing a discriminator for the frequency of the coarse frequency oscillator are avoided by providing tuning of the coarse frequency oscillator and an auxiliary oscillator by means of a common tuning capacitor shaft. The auxiliary oscillator possesses approximately a thousandth of the coarse oscillator frequency and roughly the same relative frequency variation. This oscillator comprises a three-point circuit consisting of the two transistors (T 501/T 502). The circuit arrangement provides high frequency stability due to its high Q and negligible transistor feedback. By means of DC negative feedback via R 503 optimum current distribution is induced into the differential amplifier T 502/T 503, thus ensuring a reliable oscillation commencement. The near-square wave output is brought out across the second half of the differential amplifier T 503 and the transformer Ü 501 at an amplitude of  $U_{pp} = 2$  V and drives the limiter of the following circuit.

### 2.1.6. Discriminator for Auxiliary Oscillator (6)

The capacitor combination C 603/C 604 is pulse switched either to 0 V (ground) or -15 V by the transistors T 601/T 602, according to the input signal wave form derived from the auxiliary oscillator. The capacitors C 603 and C 604 are then charged and discharged according to the amplitude swing of this square wave pulse.

The pulsating discharge current is supplied to capacitor C 605 via GI 603. Its DC component is applied to the subsequent linearizing circuit via a low-pass filter network.

To ensure that this current component is reduced with increasing frequency, it is superimposed with a constant current of opposite sense. The sum of these two currents is adjusted to 1 mA at a coarse oscillator frequency of 102 MHz by means of the potentiometer P 601. The constant opposing current is generated by T 603. GI 605 and the base-emitter diode T 603 compensate the opposing current by giving the same temperature dependence as that of the discriminator current effected by the diodes GI 603 and GI 604.

The slope of the discriminator characteristic, i.e. the change in discriminator current as a function of frequency, is proportional to the capacitance of C 603 and the charging current. The design slope of the discriminator is -0.2 mA/MHz. The additional capacitor C 604 is provided for slope alignment when working in conjunction with the linearizing circuit. The low-pass filter network on circuit board 241-AE filters out the control current required for the linearizing circuit to avoid frequency modulation by the auxiliary oscillator frequency.

### 2.1.7. Linearizing Network (7)

The coarse tuning oscillator must drive the control oscillator via the auxiliary oscillator and the following discriminator to ensure that the control oscillator frequency is 3 MHz above the coarse oscillator frequency at all times.

The task of the linearizing network is to convert the discriminator output current into the control voltage required by the control oscillator varicaps. The input current of the linearizing network is a linear function of the frequency, corresponding to the discriminator characteristic. The output voltage must be a function of the characteristic of the varicaps in the control oscillator. At point 4 of the linearizing network, the



maximum output voltage of +19 V is achieved when the control oscillator frequency is 105 MHz and the input current 1 mA. Minimum output voltage of +2.4 V occurs when the frequency of the control oscillator is 69 MHz and the input current 8.2 mA. To avoid spurious modulation of the control frequency, the output voltage of the linearizing network must be maintained highly constant. The control circuit comprising T 704 and T 705 is provided for this purpose.

The input current from point 5 is applied via L 702 to the emitter of T 707 and emerges at T 707 collector reduced by the amount of the base current. For small values of the collector current, diodes GI 706 to 714 are rendered nonconducting, resulting in the voltage at T 707 collector, and thus that at point 4 of the circuit, being determined solely by the voltage drop across R 720 and GI 705.

The voltage divider comprising R 710 to R 719 causes each of the diodes to assume a certain potential. If the collector voltage of T 707 possesses a potential with an amount less than that of the diode threshold, this will cause the respective diode to conduct.

In the absence of the diode circuit, the collector voltage of T 707 would fall as a linear function of the increasing input current. The introduction of the diode circuit, however, together with the parallel connection of an additional resistor, effects a reduction of the collector resistance at definite collector voltage values, thus causing a kink in the characteristic curve, showing the collector voltage as a function of emitter current for T 707.

The desired variations of the output voltage is achieved by adjustment at three points of the voltage divider R 710 to R 719. The initial, centre and end portions of the range are adjusted with potentiometers P 701, P 703 and P 704 respectively.

The base voltage of T 707 has been shifted with respect to 0 V by a positive amount equal to the diode threshold by means of GI 704 to ensure that the input (emitter T 707) is not DC biased with respect to 0 V. The control circuit comprising T 701 to T 703 supplies a constant voltage of - 15 V to the auxiliary discriminator and to the auxiliary oscillator itself.

This voltage also causes a constant current to flow via R 705. The control circuit output voltage is adjusted at the emitter of T 704 to compensate the above-mentioned constant current except for the small base current required by T 705. The output voltage is thus determined by the values of the resistors R 704 and P 701 and the current across

R 705. Adjustment of P 702 must always occur prior to adjusting P 701.

The zener voltage of GI 702 is increased by the forward voltage of diode GI 701 (which provides temperature compensation of the emitter-base voltage of T 703) and serves as a reference voltage for the 15 V control circuit. The control circuit compares that portion of its output voltage tapped at P 702 with the reference voltage and adjusts the output voltage correspondingly across the control P 701.

Upon switching-on, it could happen that the collector of T 701 assumes a positive voltage via R 705 thus resulting in no zener voltage, since GI 701 is then rendered non-conducting.

The circuit would then remain in this condition. The diode GI 703 and the resistor R 701 ensure, however, that this condition does not occur. A positive voltage at the collector of T 701 blocks diode GI 703. At the same time, however, current is applied to the zener diode GI 702 via R 701 which then effects a build-up of the zener voltage required to operate the circuit.

#### 2.1.8. Frequency and Phase Control Circuit (8)

This circuit has the following tasks:

- 1) Normally, to sum the control signal currents (outputs from the phase-meter and counting discriminators I and II) and to amplify this sum in a limiter amplifier prior to application to the linearizing circuit.
- 2) In the sweep mode, the phase control or the frequency and phase control circuit must be capable of being switched-off depending on the sweep width.

When the phase control circuit (Ü 801, Ü 802, GI 803 to 806) is to be switched-off, the contacts of relays rel 803 and rel 801 are opened. The transistor amplifier T 801/T 802 is then only driven by the sum of the output currents from both the counting discriminators (9) and (10). The sum of the output currents is zero under equal frequency conditions due to the opposite polarity (fed via point 1). In this case the output current at point 11 of the circuit is near zero, whereby under maximum conditions the value is in the region of  $\pm 0.8$  mA.

The frequency control circuit is switched-off by opening of contact rel. 802. This is the case when the sweep current is supplied directly to the linearizing circuit via point 3, R 808 and point 11 for the purpose of generating the four largest sweep widths

When the frequency control operates in conjunction with the phase control, the purpose of the former is to provide an accurate adjustment of the control frequency to ensure that it always remains within the locking range of the phase control. This can only be achieved by the frequency control when the characteristics of the two counting discriminators coincide to a large degree.

#### 2.1.9. Counting Discriminator I (9)

The task of this discriminator is to convert the difference frequency (control frequency minus coarse oscillator frequency) into a direct current proportional to the frequency. The slope of its characteristic amounts to an average of 1 mA/MHz; its output current is of an opposite polarity (neg. count direction) to that of counter discriminator II (see 2.1.10.). L 901/C 901 complement the low-pass filter of the frequency converter T 901 has the function of a buffer stage and T 902 an amplifier. T 903 and T 904 form an AC coupled limit amplifier, T 905 and T 906 constitute a pulse shaper, comprising two switching stages which operate alternately. These two transistors together with the shunt arrangement of C 912 and C 913 and the diodes GI 903 and GI 904 make up the frequency discriminator, the slope of which is varied by means of C 912. The discriminator operates in a similar manner to the auxiliary oscillator discriminator (6) (see 2.1.6.).

The output signal of the switching stage is applied to the phase control circuit via T 907, this, however, is only the case when point 2.1. of the circuit is switched to +36 V via the contact of relay 801 accommodated in the phase control circuit. Only then are T 907 and GI 905 rendered operative.

#### 2.1.10. Counting Discriminator II (10)

The counting discriminator II converts the output signal of the interpolation oscillator into an output current being proportional to the input frequency. Its slope corresponds to that of counting discriminator I, its output current is, however, positive thus resulting in the sum of the output currents of both discriminators being zero when the respective input frequencies are equal.

Since the square wave input signal possesses a sufficient amplitude, a pre-amplifier stage and also the AC coupling of the limit amplifier can here be dispensed with.

In all other respects the circuit arrangement corresponds to that of counting discriminator I.

#### 2.1.11. Standard Frequency Buffer (11)

The output of the standard frequency source is applied via the buffer circuit in the correct waveform to the various subunits of the three basic instruments; control oscillator, generator and receiver. The transistor stage comprising T 1101 and T 1102 amplifies and limits the input signal to a voltage of approx.  $U_{pp} = 6.5 \text{ V}$ . Alternate switching of the connecting point of both collectors to the emitter potential of either T 1101 or T 1102 is achieved by causing the corresponding transistor to conduct.

The resulting square wave signal is applied to the phase meter 14 via R 1112 and is also fed to the generator PS-5 via an attenuator, comprising resistors R 1111 and R 1113.

The frequency divider network comprising circuits (12a) and (12b) is driven via the buffer stage consisting of T 1103 and T 1104.

The circuit configuration of board 241-AW is responsible for the distribution of the 100 kHz and 10 kHz signals. This circuit also applies the current supply to the frequency divider networks.

#### 2.1.12. 10 : 1 Frequency Divider (12)

Two identical 10 : 1 frequency dividers are series connected into the instrument to generate a 10 kHz square wave signal from the 1 MHz standard. The 10 kHz signal has a dual purpose:

1. To effect frequency synchronization of the local oscillators of the receiver PM-5.
2. To generate the input signal required for the tuning indication of the interpolation oscillator.

The 10 : 1 frequency divider comprises 4 consecutive flip-flops and an output buffer. In the simple consecutive switching condition i.e. free-running, the circuit would divide the input frequency in the ratio 16 : 1. It is, therefore, necessary to disturb the normal 'binary' divider cycle to obtain the desired 10 : 1 division.

The circuit operates in such a manner that the positive drive pulse for each flip-flop is not fed directly to the flip-flop input but via diode gates. This allows the reduction of cycle from 16 to 10, which is achieved by blanking one or more driving pulses from the flip-flop and thus jumping several pulses.

The switching cycle is shown in circuit diagram (12) .

It can be seen that the first flip-flop switches quite regularly. The transistors shown to the left in the circuit diagram are conductive. This means that the diode gate GI 1201 of the first flip-flop conducts, whereas GI 1202 blocks. The first positive input pulse will block transistor T 1201 via GI 1201 and thus triggering the first flip-flop. The second input pulse via the now conducting diode gate triggers the first flip-flop back. The positive output pulse at the collector of T 1202 then triggers the second flip-flop into its operative position. The first interruption into the normal operating cycle takes place at the fourth input pulse. The second flip-flop should now be triggered back by the second positive output pulse, however, the diode gate GI 1205 is blocked because of the base emitter potential of T 1205 is held at constant -7.1 V by GI 1204. Diode gate GI 1206 however conducts and thus triggers the third flip-flop into its operative position. This means that a condition is achieved at the fourth input pulse which would normally appear firstly at the sixth input pulse. This means that two strokes have been jumped. The second interruption into the normal operating cycle takes place at the sixth input pulse, where the fourth flip-flop instead of the third flip-flop is triggered into its operative position via GI 1208. This means that four further strokes have been jumped, since this condition is achieved normally firstly at the twelfth input pulse. The rest of the cycle then continues in a regular manner. The tenth input pulse then re-commences the cycle.

The frequency division can therefore be summarized as :

The first flip-flop operates individually and forms a 2 : 1 divider.

The second, third and fourth flip-flop in conjunction with each other form a 5 : 1 divider.

The series connection of both dividers results in a 10 : 1 division. The output-buffer stage has the advantage over a normal collector-base stage by having a quicker charge reversal of C 1217. T 1208 acts as a high-impedance emitter resistance if T 1210 is triggered by a negative voltage step of approx. -1 V to -7 V applied to its

base and if its emitter follows this voltage. T 1210 is initially blocked due to C 1217 charging-up during subsequent reversed voltage gradient. Charging reversal is quickly effected, however, since T 1208 commences to conduct.

### 2.1.13. Standard Frequency Oscillator (13)

This unit comprises:

- a) a 2 MHz crystal oscillator with a 2 : 1 frequency divider and a stage designed to switch-off the oscillator when an external standard frequency is used.
- b) a two-stage limiter for the output signal.

This oscillator supplies an internal 1 MHz standard frequency, the accuracy of which determines the frequency accuracy of the measuring set.

This oscillator, comprising transistors T 1301/T 1302, operates as a Colpitt's oscillator. The crystal is not operated at its series resonance but between the series and parallel resonance and thus acts as an inductance. Coupling of the two transistors T 1301 and T 1302 via a common emitter resistor fulfills a dual purpose:

1. Low input capacitance over the whole operating range thus causing the circuit to have little effect on the frequency stability and the simultaneous amplitude limiting.
2. Output voltage free of retro-action.

Use is made of the following two trimmers to effect frequency adjustment in the order of  $\pm 50 \times 10^{-6}$  to  $\pm 80 \times 10^{-6}$ :

$$C 1301 : \cong \pm 35 \times 10^{-6}$$

$$C 1305 : \cong \pm 15 \times 10^{-6}$$

Drift due to ageing, etc. can thus be compensated for. Initial drift due to aging amounts to approx.  $1 \times 10^{-7}$  per week.

The crystal oscillator is followed by a 2 : 1 frequency divider in the form of a flip-flop (transistors T 1305/T 1307). T 1304 supplies the required square-wave signal having a steep leading edge.

The frequency divider output controls a two-stage limiter. The stages, which operate as differential amplifier, are emitter-coupled with respect to AC voltages only, thus enabling a stable DC operating point to be achieved inspite of the tolerances of the individual transistor characteristics. The output of T 1309 is applied to the internal

subunits via socket Bu 1302 at an output impedance of  $Z_{out} = 75 \Omega$ . T 1310 delivers its DC free output also at an output impedance of  $Z_{out} = 75 \Omega$  to socket Bu 1303 for connection of external loads.

If the frequency is to be controlled by an external standard frequency applied to socket Bu 1301 instead of the internal crystal oscillator, transistors T 1303/T 1306 will cease to supply DC to the 2 MHz oscillator if the external frequency standard has a DC resistance of  $\leq 100 \text{ k}\Omega$ . The limiter receives its input signal then from T 1308.

#### 2.1.14. Phase Meter (14)

The purpose of this circuit is to generate a control voltage for effecting the frequency control of the coarse frequency oscillator. This circuit also supplies signals to the respective meter amplifier (15) to indicate the phase-locked condition. The unit (14) contains a 1 MHz and a 100 kHz phase comparison circuit. The 1 MHz square wave signal is applied to point 2 of the 1 MHz phase comparison circuit, from which a fast narrow switching pulse is formed by the pulse former T 1401/T 1403 for the diode chain comprising G1 1403 to G1 1406. Under quiescent conditions transistors T 1401 and T 1403 are blocked. The input square wave voltage is differentiated via C 1403/R 1408 and causes T 1403 to conduct. The differentiation of the negative output pulse from T 1403 is achieved by the variable capacitance diode G1 1402. Since this diode was turned on by the bias current via R 1411, the capacitance of G1 1402 required for differentiation is effective only after being sufficiently discharged. The width of the resulting pulse is thus mainly dependent on the discharge time and can be adjusted by means of the bias current. The pulse resulting at transformer  $\ddot{U}$  1401 switches on all of the diodes G1 1403 to G1 1406. The instantaneous voltage present at the emitter of T 1404 during conduction (coarse frequency oscillator signal via buffer stage T 1402, T 1404) is thus applied through the low-pass filter R 1414/C 1406 or R 1415/C 1406 to the output at point 5. The phasing of the oscillator signal with respect to the switching pulse determines the output voltage wave-form .

Under constant phase conditions and if the oscillator is phase-locked, a positive or negative DC voltage is generated. If the oscillator is not phase-locked, an AC voltage results. The frequency components of this voltage consist of the difference frequencies between the coarse oscillator frequency and all multiples of 1 MHz.

In order to obtain an output signal with an amplitude independent of the oscillator frequency, it is necessary for the conducting period of the diodes to be shorter than one half cycle of the highest oscillator frequency (102 MHz). If the switch-on time corresponded to one cycle of the oscillator frequency, no output signal would result. During the pulse intervals under 'worst case' conditions, the diodes would cut off the peak-to-peak oscillator voltage at the emitter of T 1404. This would result from the threshold of the series-connected diodes 1 N 914 (GI 1403/GI 1404 or GI 1405/GI 1406). The series arrangement of R 1414 and R 1415 presents a load to the transformer and hinders the formation of interference transients by the transformer which would otherwise occur under no-load during the pulse intervals.

The 100 kHz phase comparison circuit has the following differences with respect to the circuit previously described:

The output voltage resulting from the phase comparison is superimposed on a bias voltage of -10 V. This voltage is presented to the circuit at point 6 from the zener diode GI 402 in the coarse frequency oscillator.

Positive feed-back through C 1413 to the pulse shaper input is provided to steepen the leading edge of the output pulse. R 1424 prevents the output impedance of the controlling 10 : 1 frequency divider (12a) from affecting the feed-back.

The voltage resulting at point 7 of the circuit sets the coarse tuning oscillator frequency to the desired value by means of the variable capacitance diode GI 401 in the coarse tuning oscillator.

#### 2.1.15. Lock-In Indicating Amplifier (15)

This circuit receives the phase-meter output signals and indicates, on meter J 2302, whether the coarse frequency oscillator is locked and whether its frequency is a multiple of 1 MHz or 100 kHz.

As with the phase-meter it is useful to discriminate between a 100 kHz channel (input at point 1) and a 1 MHz channel (input at point 5) when discussing the lock-in indicating amplifier. A discriminator is connected to the 100 kHz channel output, whereas a rectifier circuit for the meter J 2302 is present at the output of the 1 MHz channel.

In the phase-locked mode no AC signal is present at the 100 kHz channel input and thus the discriminator circuit output does not supply DC to the meter circuit.



The result of this is that the phase-locked mode is indicated. In the absence of a DC output in the 1 MHz channel (the rectifier circuit has no DC output when the coarse frequency oscillator frequency is a multiple of 1 MHz), the meter indicates the 1 MHz marker. For all other phase-locked frequencies the 1 MHz channel has a constant DC output resulting in the meter indicating the 100 kHz marker, meaning that the adjusted and phase-locked frequency results in an integral multiple of 100 kHz but not of 1 MHz.

The input of the 100 kHz channel contains a mixture of all the difference frequencies between the coarse frequency oscillator frequency and the neighbouring 100 kHz spectrum lines. After being amplified by T 1501 the signal is applied through the matching stage T 1502, and a low-pass filter having a cut-off frequency of 55 kHz and poles at 100 and 150 kHz, is amplified once again (T 1503, T 1504) and then applied to the pulse shaper stage T 1505 of a counting discriminator. Whenever transistor T 1506 is blocked, C 1510 can discharge into the zener voltage of GI 1504 via GI 1505 and R 1510. The mean charging current flows through meter J 2302 and is proportional to the switching frequency of T 1506.

At 50 kHz the meter current amounts to 700  $\mu$ A corresponding to 1.7 times the full scale value if transistor T 1505 fully conducts. At the limit of the lock-in range, the meter current is reduced to approx. 28  $\mu$ A (7 % of the current at f.s.d.) and thus aids in locating the lock-in range and recognizing the phase-locked condition during tuning. When the lock-in range is attained, the meter needle swings into the 1 MHz section of the meter scale and, as long as the pull-in range is not exceeded, remains steady despite the coarse frequency oscillator setting being changed.

This phase-locked condition is so obvious that a dip in the tuning indicator reading between two lock-in frequency indications can easily be distinguished.

If the coarse tuning oscillator frequency lies between two lines of the 100 kHz spectrum, the resulting 50 kHz difference frequencies can cancel each other out. Very careful adjustment of the coarse tuning oscillator will enable this position to be located at the upper band limit.

This occurrence will also be apparent at the input of the 1 MHz channel when locking the frequency of the coarse frequency oscillator to a multiple of 500 kHz.

Due to the constancy of the locked frequency, the two 500 kHz signals cancel

each other out and an apparent 1 MHz locked frequency indication results. The input of the 1 MHz channel contains a mixture of all the difference frequencies between the coarse tuning oscillator frequency and the neighbouring 1 MHz spectrum lines. After being amplified by T 1510 the signal is applied through the matching stage T 1509 and a low-pass filter having a cut-off frequency of 550 kHz, is amplified once again by T 1508 and then applied to the limiter comprising T 1506/T 1507. The square-wave output current of the latter is rectified to an average value representing a constant DC for meter J 2302, this current being widely independent of frequency and input voltage.

#### 2.1.16. Saw-Tooth Generator (16)

The circuit (16) supplies voltages required for sweeping the oscillator in the control oscillator section RWO-5, for the horizontal deflection of the display unit and the fly-back pulse during the saw-tooth trailing edge period.

The sweep frequency can be adjusted in steps between 0.1 Hz and 50 Hz. The peak sweep voltage amplitude is  $\pm 3.5$  V.

The capacitor selected with switch S 1601/11/b is charged with a constant current through R 1601. To ensure the linearity of the increasing voltage, the capacitor must be only subjected to high-impedance loading. Three series-connected collector stages including T 1601, T 1603, T 1604 and T 1605 constitute a low-impedance, saw-tooth voltage source. The output voltage at the emitter of T 1604 is superimposed by the voltage of the zener diode G1 1602, the cathode of which is connected to the charging resistor R 1601. Since the charging voltage is constantly increased by the amount of the capacitor voltage, the voltage across R 1601 and thus the charging current remains constant (bootstrap-circuit).

The capacitor voltage also represents the collector voltage of T 1606. When the positive maximum value is attained, T 1606 conducts and discharges the capacitor down to its maximum negative value. Direct use of the saw-tooth voltage is made to control the switching. During voltage increase, transistors T 1611 and T 1608 are initially blocked until a value is attained ( $+ 3.5$  V) at which T 1611 conducts. The current through T 1610 is reduced by the common emitter resistance and as a result T 1609 is blocked. Due to the absence of the T 1609 collector current, the base voltages of T 1607 and

T 1610 fall to a negative value determined by the voltage divider comprising R 1624/R 1625 and P 1603. The negative voltage step causes T 1606 and T 1607 to conduct simultaneously, thus initiating the discharge of the capacitor. The discharge current is determined by R 1617. The capacitor voltage falls to a value ( $-3.5$  V) at which T 1611 is blocked, resulting in all of the following transistors being switched once again. Due to the amplification offered by the feed-back loop between T 1610 collector and base, switching from charge to discharge occurs spontaneously. The positive and negative thresholds, at which switching is to occur, is adjusted with P 1602 and P 1603 respectively. During fly-back, T 1608 is also rendered conducting. The positive pulse at the collector of the latter is amplified also by T 1602 and is available at point 2 of socket Bu 1601 as a low-impedance signal for driving the display unit plug-in SGE-10.

The Display Unit SG-1 and its plug-in SGE-10 extend the Level Measuring Set PSM-5 to constitute the Sweep Measuring Set WM-50. In position 1 of switch S 1601, the deflection voltage can be manually adjusted with P 1601 for static measurements.

The peak amplitude is the same as for sweep measurements  $\pm 3.5$  V.

In position 11, provision is made for introducing an external voltage, also of  $\pm 3.5$  V peak amplitude, for deflection purposes at socket Bu 1601, point 5.

When a sweep frequency lower than that selectable is required, switch S 1601 can be switched to position 10 and an additional capacitor connected parallel with C 1611 across socket Bu 1601 at point 4, thus enabling any sweep frequency below 50 Hz to be obtained. The upper limit of the additional capacitance is determined by the leakage current which must not exceed approx.  $2 \mu\text{A}$ . Sweep width adjustment is achieved with switch S 1602. This switch switches the dropping resistors R 1629 to R 1644 across which the sweep current is supplied to the individual circuits. The sweep currents are determined by dropping resistors from the constant, peak saw-tooth amplitude of  $\pm 3.5$  V. In positions 3 to 12 of switch S 1602, the interpolation oscillator is modulated by sweep currents of  $\pm 35 \mu\text{A}$  to  $\pm 35$  mA.

In positions 13 to 17 of switch S 1602, the frequency and phase control stage is modulated by sweep currents of  $\pm 35 \mu\text{A}$  to  $\pm 700 \mu\text{A}$ .

In positions 18 to 21 for large sweep widths, the sweep currents are applied to the linearizing network. Here the currents range from  $\pm 385 \mu\text{A}$  to  $\pm 35 \text{ mA}$ .

#### 2.1.17. 2.7 MHz Crystal Oscillator (17)

This oscillator generates the auxiliary frequency of 2.7 MHz required to convert the output frequency of the interpolation oscillator to between 2.99 and 3.1 MHz. The crystal oscillator is phase-locked to the standard frequency. This local oscillator comprises the common-emitter limiter T 1706/T 1707. The oscillator voltage is created in the 2.7 MHz tuned circuit in the collector lead of T 1706, the amplitude of which is determined by R 1716 and the emitter current of T 1706/T 1707. Feed-back is achieved from the winding 3-4 of the transformer  $\ddot{U}$  1702 via the series configuration of capacitance diode G1 1703 with the resonant crystal Q 1701. This series circuit together with R 1718 presents a resistive divider for the oscillator frequency, which causes only a small reduction of the feed-back voltage for the desired main resonant frequency. For the neighbouring resonant frequencies of 2.76 and 2.79 MHz, however, the dividing ratio is so unfavourable that the amplitude condition for initial oscillation is no longer given. To ensure initial oscillation for transistor combinations having extreme characteristics, the operating point T 1706 is stabilized by a negative DC feed-back through R 1713. Otherwise it could happen that T 1707 takes over the full current on conducting thus resulting in loss of initial oscillation. The 2.7 MHz signal is decoupled via the collector of T 1707.

With the aid of the capacitance diode G1 1703, the oscillator frequency can be influenced by the control voltage. A coincidence circuit comprising transistors T 1702, T 1703 supplies this control voltage by comparing the phase of a 100 kHz signal derived from the standard frequency with that of the oscillator output signal. T 1704 acting as a blocking oscillator generates an output pulse of approx. 150 ns across coil 5-6 of the transformer  $\ddot{U}$  1701, this pulse being phase-locked to the 100 kHz standard frequency signal. The transistors T 1702/T 1703 conduct during the pulse duration. Current distribution through the transistors is determined by the phase condition of the 2.7 MHz signal. The output is obtained directly at  $\ddot{U}$  1702 via the buffer stage T 1705. If, for instance, the base of T 1703 remains negative during conduction of the coincidence stage, T 1702 will always assume the current, i.e. output current

pulses result at the collector of T 1702 which are integrated by the following low-pass circuit comprising C 1702, R 1705, R 1710 and C 1706 and generate a negative control voltage of approx. -12.5 V.

In the other case, when positive voltage is present at the base of T 1703 during conduction, no voltage will result. Between these two limits, any voltage value is possible depending on the phase condition. In the absence of synchronization, an AC voltage will result whose frequency corresponds to the difference between the oscillator frequency and the desired frequency.

The blocking oscillator is driven via the buffer stage T 1701. The square wave output of the latter is differentiated by C 1704 and R 1704 to generate positive spikes for triggering the blocking oscillator.

Blocking oscillator conduction is determined merely by the inductance offered by winding 3,4 of transformer Ü 1701 and the value of the supply voltage.

#### 2.1.18. Interpolation Oscillator (18)

This oscillator serves to effect fine adjustment of the control frequency. Strict requirements are thus made on its stability. With the aid of a reactance circuit acting in parallel with the oscillator tank circuit, sweep deviations up to 17.5 kHz can be produced with the aid of the output signal of the saw-tooth generator.

The transistors T 1805 and T 1806 form an emitter-coupled differential amplifier. The output current of T 1806 as determined by R 1816 and base voltage bias is applied to the resonant circuit mounted on printed circuit board 241-C. Feed-back is achieved via the base of T 1805. The oscillator amplitude is governed by resistor R 1817 and it is this resistor that determines the load impedance of T 1806. The output from the common-base stage T 1804 has an exceptionally low retro-action.

The complete circuit, including the reactance circuit comprising T 1807/T 1808, is operated with a stabilized, filtered DC voltage, since otherwise interference voltages superimposed on the supply voltage would cause spurious modulation of the reactance circuit.

The reactance circuit comprising transistors T 1807/T 1808 forms an inductance in parallel with the resonant-circuit inductance L 1803; the former reducing the latter by approx. 3 %.

The inductance is proportional to the values of C 1814 and P 1801, R 1818 and inversely proportional to the current flowing through T 1807/T 1808. This current is modulated to effect frequency modulation. The saw-tooth current modulated in the emitter of T 1809 results in a proportional control voltage across R 1826, which is present at the base of T 1808 and due to the high emitter resistance of this transistor, generates a proportional current variation.

The deviation of  $\pm 17.5$  kHz, which should result at a saw-tooth current of  $\pm 3.5$  mA, is adjusted with potentiometer P 1801. The selected deviation must be independent of the selected oscillator frequency. In the absence of a correction circuit, however, the deviation would vary according to the oscillator frequency range 2.9 to 4 MHz in the ratio 4/2.9. For this reason the inductance L 1802 is placed in parallel to capacitor C 1814 thus resulting in an increase of circuit capacitance with increasing frequency, thus opposing the effect of the reactance circuit. In this manner a near constant deviation is achieved over the whole frequency range.

The oscillator output signal is decoupled across R 1815 and controls, via T 1803, the limiter amplifier comprising T 1801/T 1802. The output current of T 1801 feeds the tuning indication (19) and the 10 : 1 frequency divider (22).

The presence of a signal at socket Bu 1801 is determined by the condition of diode G1 1801. When a voltage of + 6.3 V is available at point 2, the diode conducts and the output signal is short-circuited by C 1801. This is desired when the counting discriminator II (10) is to be driven with the interpolation oscillator frequency converted in the range 2.99 to 3.1 MHz (interpolation range 110 kHz).

#### 2.1.19. Tuning Indication Circuit (19)

This circuit and the subsequent discriminator have the task of indicating the exact frequency tuning of the interpolation oscillator in increments of 10 kHz at meter J 2301.

To do this, the circuit receives a 10 kHz signal from the 10 : 1 frequency divider (12b) which is derived from the standard frequency oscillator. This signal is converted into narrow pulses, having a repetition frequency of 10 kHz, by transistors T 1902 and T 1906. T 1902 operates here as a buffer stage and generates by means of differentiation (C 1904, R 1912) control pulses for the blocking oscillator T 1906. The pulse

During the duration of the pulse the blocking oscillator output causes the differential amplifier comprising transistors T 1904 and T 1905 to conduct. The coincidence of this conduction period with the input signal at T 1904, which is derived from the interpolation oscillator, controls the magnitude of the output pulse from T 1905. The circuit thus operates, in principle, as a phase comparator.

The collector current pulses from T 1905 are integrated by the 5 kHz low-pass filter resulting in the difference frequencies of the interpolation frequency being available at the low-pass filter output for the following 10 kHz increments. An amplifier following the low-pass filter supplies the subsequent discriminator with the required control voltage.

## 2.1 ). Frequency Converter (20) , Buffer and Switching Stage (21)

The purpose of the frequency converter is to convert the interpolation frequency 290 to 400 kHz (divided 10 : 1 by the frequency divider (22) ) into the frequency range 2.99 to 3.1 MHz.

The switching stage has the task of supplying the counting discriminator II with either the 2.9 to 4 MHz signal supplied directly from the interpolation oscillator or the converted signal.

A low-pass filter mounted on circuit board 241-D at the signal output of the mixer allows only the fundamental of the 290 to 400 kHz square-wave input signal to pass. The mixer itself consists of the two transistors T 2001 and T 2002 and operates as a mixer in push-pull configuration to suppress the carrier. An ideal push-pull mixer chops the marker signal voltage as dictated by the carrier frequency and thus acts as a switch. This is achieved here by a differential amplifier, which is turned on and off at intervals dictated by the carrier frequency. The common-emitter impedance of the amplifier consists of the parallel arrangement of the high-impedance of the carrier current source and resistor R 2008. The task of this resistor is to ensure a rapid discharge of the stored charge of the carrier transistor during blocked conditions. The large negative feedback resistors in the emitter leads result in a high linearity and balance.

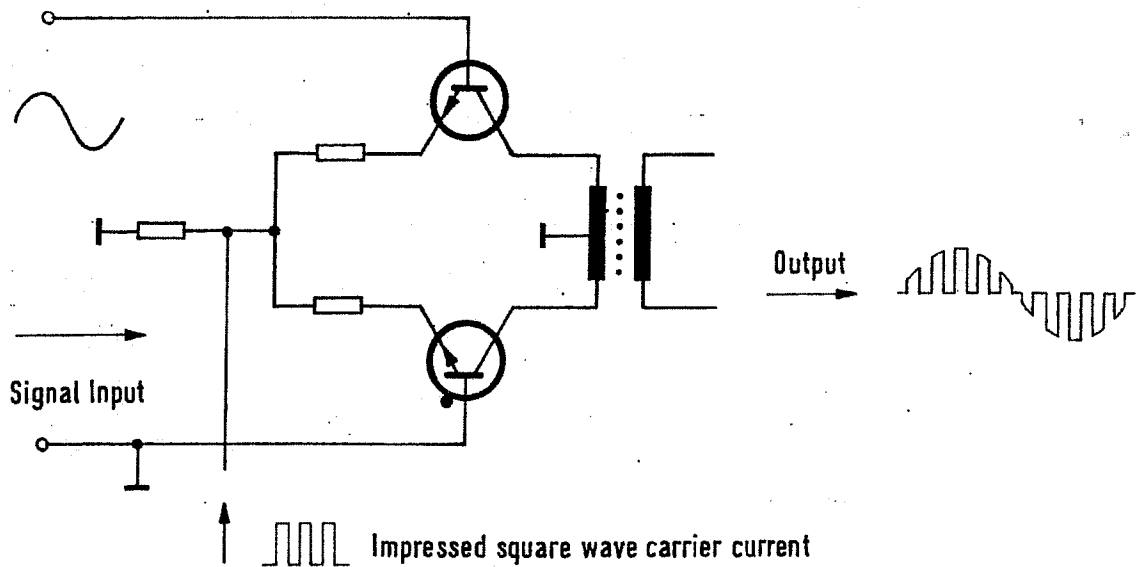


Fig. 3

The sum frequencies 2.99 to 3.1 MHz occurring following the mixer are selected by a sharp band-pass filter arrangement, which possesses a particularly high attenuation for the carrier frequency of 2.7 MHz and the difference frequencies of 2.3 to 2.41 MHz.

Compared with simple diode modulators operating in push-pull, the above mentioned circuit has the advantage that the carrier and marker signals can be applied unbalanced and also that the attenuation caused by the mixer can be compensated by simultaneous amplification.

Following the band-pass filter at the mixer output, an amplifier is provided which drives the emitter-coupled frequency limiter stage comprising T 2104/T 2103. When a voltage of + 6.3 V is present between point 2 of circuit board 241-J and ground, diodes GI 2106 and GI 2101 conduct whilst GI 2102 and GI 2103 are blocked. This condition of the switching stage (gate circuit) is indicated by pilot lamp 2101. The output signal from T 2103 is short-circuited by GI 2101 and separated from resistor R 2115 by GI 2102 and GI 2103. Since the voltage of + 6.3 V (measured against ground) at point 2 of the interpolation oscillator (circuit board 241-A) is simultaneously turned-off, the output of the interpolation oscillator is now applied to diodes GI 2105 and GI 2104, which are no longer blocked, to the resistor R 2115 and is thus also applied to the input of counting discriminator II.



### 2.1.21 10 : 1 Frequency Divider (22)

The circuit divides the frequency of the interpolation oscillator and supplies the signal to the frequency converter (20) which then adds 2.7 MHz. The output frequency of 2.99 to 3.1 MHz following the frequency converter is thus varied by only 100 kHz when the frequency of the interpolation oscillator is varied by 1 MHz. This enables an expansion of the interpolation oscillator scale by a factor of 10 to be achieved.

The low-pass filter located at the frequency converter output has the task of selecting the fundamental. This is assisted by arranging for the divider output signal to have a duty cycle of 1 : 1 thus suppressing the second harmonic.

It is for this reason that the frequency dividers (22) and (12) are each of a different design.

Divider (22) comprises a 5 : 1 divider working into a 2 : 1 divider thus enabling the desired duty cycle of 1 : 1 to be achieved. The table shown in circuit diagram (22) illustrates the relative changes in the operating cycle of the transistors and diodes.

### 2.1.22. Discriminator (23)

This discriminator converts the output of the tuning meter circuit (19) into a direct current proportional to this frequency for meter J 2301. The circuit arrangement of the tuning indicator-discriminator-meter circuit enables the interpolation oscillator to be easily compared with a 10 kHz spectrum line of the standard frequency at an accuracy of better than 20 Hz.

An almost sinusoidal control voltage with a peak-to-peak amplitude of 2V is applied to the discriminator input causing the alternative conduction of transistors T 2301 and T 2302. This results in a square-wave voltage of approx. 11 V at C 2304, which charges and discharges the capacitor as dictated by the input frequency. The charging current surges are rectified and generate a DC voltage across C 2305. A meter current proportional to this DC voltage flows through the branch comprising R 2304 and meter J 2301 which are both in parallel with C 2305. The discriminator characteristic is non-linear since the DC across C 2305 increases with increasing frequency and thus the amplitude swing, which serves to charge C 2304, is reduced. The slope of the characteristic is thus reduced with increasing frequency.

In the sweep mode the false impression must not be given that the centre frequency of the interpolation oscillator can be exactly adjusted. To enable this meter 2301 receives an additional current which generates a full-scale deflection during sweep operation. In all positions of the sweep deviation switch S 1602 the + 6.3 V voltage at point 5 on circuit board 241-AJ is switched off thus rendering GI 2306 non-conducting. This results in a voltage of - 9 V across GI 2306 and generates sufficient current through R 2311 for a full-scale deflection.

Under normal operating conditions approx. + 0.65 V is present across GI 2306. The effect of this voltage upon the meter reading is compensated for by resistor R 2309. GI 2305 limits the current through J 2301 to approx twice or three-times f.s.d. Circuit board 241-AJ also includes a relay which is activated in the positions 18, 19, 20 and 21 of the mode selector switch 1602, thus enabling an additional current to be applied to meter J 2302 for f.s.d. This is to avoid an apparent phase-lock indication of the coarse tuning oscillator at sweep deviations in excess of  $\pm 1.75$  MHz.

The emitter stage comprising transistor ASY 26 is accommodated on circuit board 241-C for reasons of space and serves to filter an auxiliary voltage for the interpolation oscillator. Interference voltages superimposed on this voltage would lead to an undesirable frequency modulation of the oscillator.

### 2.1.23. Power Supply (24)

The power supply has the following tasks:

To supply the instrument with 5 different DC voltages each independent of mains voltage fluctuations.

To provide both AC line and battery operation

This unit comprises the following circuits:

- 1) Voltage supply for two-point regulator
- 2) Two-point regulator
- 3) Chopper
- 4) Rectifier and filter network following 3) above

This power supply features high efficiency and compact dimensions.

## 1) Voltage Supply for Two-Point Regulator

In the AC line mode, the mains AC is applied via the power transformer Ü 2401 (suitable adjusted) to the bridge rectifier circuit comprising diodes GI 2403 to GI 2406. The chokes L 2402 and L 2403 reject high-frequency interference superimposed on the AC line voltage.

In the battery operating mode, the direct voltage is applied through socket Bu 2401, choke L 2401 and the protective diode GI 2401 to the parallel arrangement of the charging capacitors C 2401/C 2402. Whereby battery ground polarity is arbitrary. Choke L 2401 protects diode GI 2401 from excessive transient currents, the latter ensuring that an incorrect battery polarity connection will cause no damage. Should both battery and AC line supply voltage be switched on simultaneously, loading is effected only on the DC source when the rectified AC (voltage across the charging capacitor) falls to a value below that of the battery voltage.

## 2) Two-Point Regulator

The purpose of this circuit is to maintain the mean value of the voltage across capacitor C 2404 to approx. 19 V and to limit the charging current for this capacitor to a value dictated by P 2401 of 6.5 A.

Transistor T 2401 then assumes the function of an ON/OFF switch; the diode GI 2407 receiving the current during the blocking period of this transistor.

The most important part of the circuit is the Schmitt-trigger comprising transistors T 2403, T 2405 and T 2406 which is shown in the following diagram:

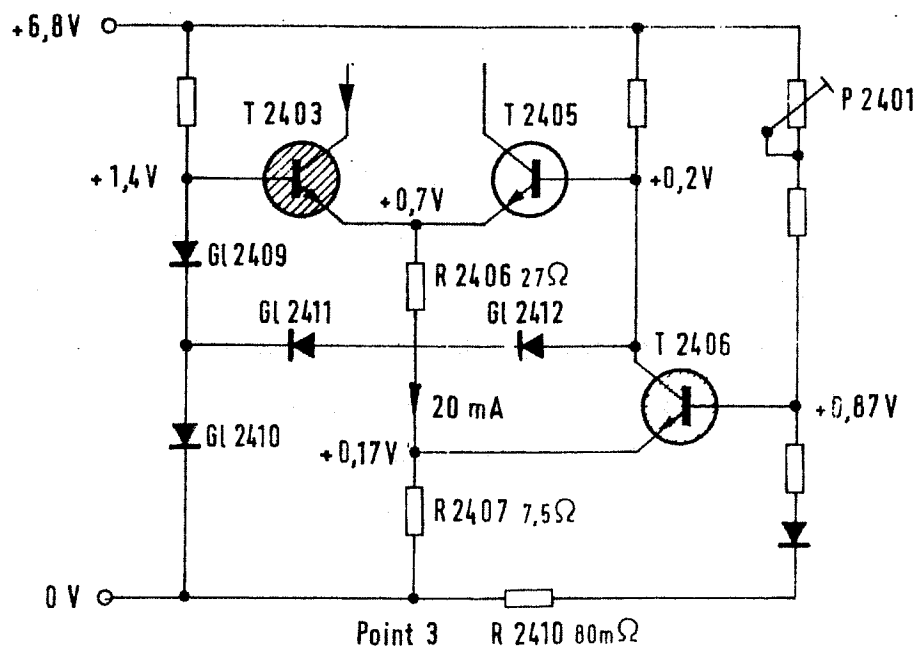


Fig. 4

T 2403 is conducting (indicated by the shading). The voltage at the base of T 2406 must fall below + 0.87 V to actuate the trigger stage.

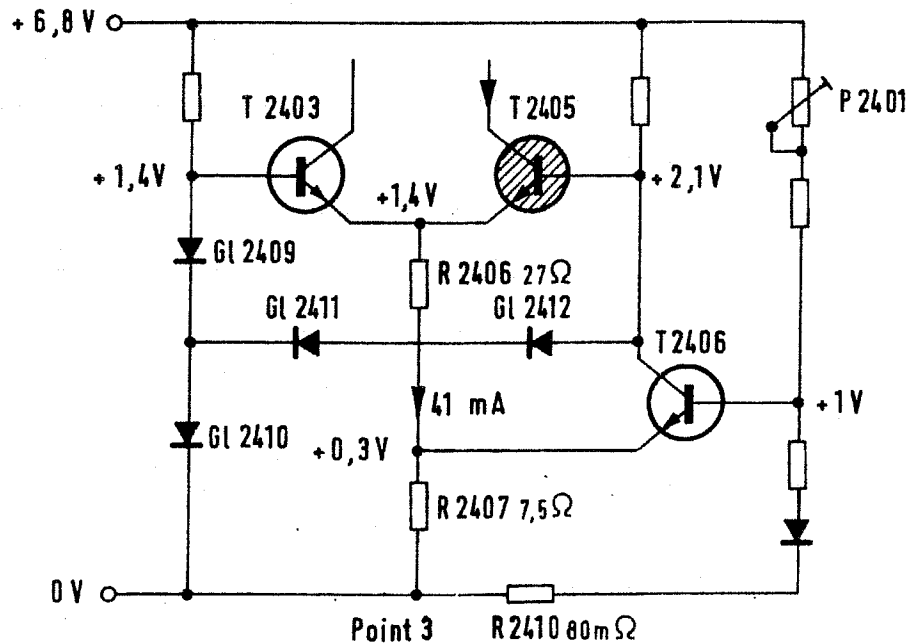


Fig. 5

T 2403 is now blocked. The triggering stage switches as soon as a voltage of + 1 V is exceeded at the base of T 2406.

All potential values given in the circuit diagram are referenced to 0 V (point 3 of printed circuit board 241-N).

All operative potentials of the circuit are determined by the forward voltage of diodes G1 2409 to G1 2412 together with the emitter-base voltages of transistors T 2403 and T 2405 and the voltage drop across R 2406. When T 2403 is conducting a voltage of + 1.4 V (two silicon threshold voltages in series) is formed at the base and a voltage of + 0.7 V at the emitter. When T 2403 is blocked (T 2405 conducting) a voltage of + 1.4 V can be measured between the common-emitter and ground. The currents are therefore determined by the resistors R 2406 and R 2407 together with the threshold ratio of the trigger.

If T 2405 conducts, because the base voltage of T 2406 is less than + 0.87 V, T 2402 will receive a base current of 41 mA, which is sufficient to cause itself and T 2401 to fully conduct. C 2403 is then charged via G1 2402 to 0.3 V. This charge is then available for the discharge of the charge carrier via R 2404 on blocking T 2401 if

T 2402 is blocked. The base voltage of T 2406 must exceed + 1 V if T 2401 is to be blocked. T 2403 then conducts followed by T 2404 which causes, practically speaking a short circuit between points 4 and 6 of the printed circuit board 241-N. However T 2402 and T 2401 will be blocked because of this.

The base voltage of T 2406 is governed by 2 different magnitudes:

Firstly the voltage drop across R 2410 and thus the charge current of capacitor C 2404 and secondly by the output signal of T 2407. If one does not consider this output signal, because eg. T 2407 is blocked, then the charge current of C 2404 will increase, after T 2401 conducts, until the voltage drop across R 2410 and the auxiliary current through P 2401 are sufficient to overcome the upper trigger threshold. T 2401 will then be blocked until the lower trigger threshold is overcome due to the reduction of the charge current. The charge current thus varies continuously to and fro between two values, which are, when T 2407 is blocked, only dependent of the position of control P 2401.

T 2407 is blocked during short-circuit of C 2404. In this case P 2401 is adjusted so that the short circuit current flowing amounts to 6.5 A. Since the trigger thresholds determine the maximum values of the current through R 2412, the threshold ratio can be easily found by measuring the AC voltage across R 2427...2430, the peak-to-peak value being identical to the delta voltage displayed at the oscilloscope.

The output voltage of T 2407 will become effective if the DC voltage across C 2404 deviates from an adjustable nominal value. One should really observe the actual value across C 2404. It is, however, advantageous to measure the real value before L 2405 at points 2 and 5 of the printed circuit board 241-N, since the phase shift of L 2405/C 2404 is avoided and the control stability is easier to achieve. It is, of course, desirable since only the DC component is to be kept constant and a constant load may be assumed, this method is permissible. The voltage taken from points 5 and 2 is divided (R 2418, R 2420, R 2419, R 2402, R 2425), averaged over approx. 2 ms (C 2406) and referenced, via T 2408 with the zener voltage at Gi 2414. The output signal of T 2408 is amplified by T 2407 and has an effect on the voltage at the base of T 2406. The output voltage of the two-point regulator can be adjusted with P 2402.

The two-point regulator switches at a frequency of approx. 1 kHz to 35 kHz according to the input voltage. The on and off switching moments could be synchronized by an interference voltage or feed-back from the following choppers. Ripple voltages at 100 Hz would appear during synchronization at the power supply output during AC

operation. To avoid this the choke L 2406 and capacitor C 2407 have been provided in addition to the wiring and screening measures.

### 3) Chopper

The chopper has the task of forming the various operating voltages for the unit from the output voltage of the two-point action control. The chopper generates square-wave voltages at the output of transformer Ü 2402 at a frequency of 26 to 30 kHz. In contrast with the usual circuitry, which utilize the saturation of the chopper transformer, the chopper transistors are clearly switched by a control current in the base circuit in order to achieve minimum loss.

In the description of the circuit, it is assumed that T 2409 is conducting and T 2410 is blocked. The base current of T 2409 is constant at this moment. The voltage for this is supplied by the feed-back winding 5-6. When T 2409 commences conducting, the end of winding 7 is positive in respect to the centre-tap 8. The voltage drop on the collector-emitter path is practically zero and the constant battery voltage of approx. 19 V is available across winding 7-8. A constant voltage is now also induced into the feed-back winding 5-6 which, corresponding to the turns ratio, will amount to approx. 9 V. The constant base current flows from the positive end 5 of winding 5-6 through resistor R 2422, GI 2416, the emitter-base path of T 2409, R 2421 and returns back to winding 5-6. This base current is so large that transistor T 2409 is saturated even at maximum collector current and minimum current amplification.

A further oppositely orientated current now flows from the positive end 3 of the second feed-back winding 3-4 through choke L 2407, GI 2417, base-emitter path of T 2409 and GI 2419 and returns to winding 3-4. This current commences at a very low level and does not allow the base current of T 2409 to be materially less than the current through R 2421. The opposing current increases, however, with time and thereby causing a corresponding reduction of the base current of T 2409.

The switching cycle is actuated by reversal of the voltage at the windings, if the base current sinks to below the minimum value necessary to maintain the collector current. This can, however, occur only when the charge carriers of the base-emitter path are removed, which is also achieved with current via L 2407.

The base current is therefore firstly zero and increases afterwards in an opposite direction until the charge carriers are cancelled out.

The following voltage reversal at the windings causes the complete blocking of T 2409 and allows T 2410 to conduct. The cycle is now repeated for T 2410.

Without resistor R 2423 the chopper would not commence operation on applying the voltages, because both transistors would be firstly blocked. Both transistors are, however, supplied with base current so that the transistor with the greater amplification will be switched by the commencing feed-back. This "start" current complements the transient base current during chopper operation. Resistor R 2426 and capacitor C 2423 serve in the suppression of any interfering resonances caused by the winding capacity and stray inductance of the chopper transformer. A pure square-wave without overshoot is thus obtained and suppresses interference, such as oscillation at another frequency than that determined by L 2407.

#### 4) Rectification and filtering circuitry following the chopper

These circuits do not possess any special features. Because all voltages are tied together, only one voltage can be set to any specific value by means of the output voltage of the two-point regulator. Stringent requirements have been made in regard to a hum and noise-free heater supply for the receiver. The 6.3 V DC heater voltage chosen fulfils these requirements.

The voltage stabilizer T 2501, T 2502, T 2503 (circuit board 241-AX 1) is provided for the individual adjustment of the 2.9 V operating voltage of the gas pressure relays in the receiver portion.

## 2.2. Generator Portion PS-5

Operation of the complete unit:

The Generator portion PS-5 is always a part of a combination. It is either combined with the Oscillator RWO-5 to form the complete Generator PSO-5 or in conjunction with the Oscillator RWO-5 and the Receiver portion PM-5 to form the Level Measuring Set PSM-5.

The generator supplies an r.m.s. output voltage of approx. 50  $\mu$ V to 1 V. The output impedance is adjustable from 50 to 150  $\Omega$  to match different load impedances. The frequency range of the generator covers the band 10 kHz to 36 MHz. A built-in, disconnectable amplitude control ensures that the indicated output voltage remains constant. The calibration of the output voltage is suitable for the various conditions encountered:

Selection can be made between voltage level and power level calibration for the six output impedances provided.

The generator frequency of 10 kHz to 36 MHz is formed as the difference between the RWO-5 control frequency of 69.01 to 105 MHz and the fixed frequency of 69 MHz. The 69 MHz signal frequency is generated within the generator portion and is phase-locked to the 1 MHz standard signal of the RWO-5. The control signal from the RWO-5 is fed via a limiter to the frequency converter in the PS-5 as the "carrier frequency".

The generator frequency characteristics are therefore exclusively determined by the Oscillator portion RWO-5.

The difference frequency, ( $f_{\text{carrier}} - f_{\text{signal}}$ ) at the output of the frequency convert is fed through a low-pass filter, to suppress the sum frequency and amplified to the required amplitude. The final amplifier possesses an output impedance of 50  $\Omega$  and works into an output attenuator with  $Z = 50 \Omega$ . The various output impedances are formed by additional series resistors subsequent to the attenuator.

The circuitry for the output voltage indication operates as a mean value rectifier and monitors the original voltage preceding the output impedance network of the final amplifier. The indication is only valid during  $Z_{\text{out}} = Z_L$  conditions.



The amplitude control forms, together with the indicator circuitry, two control stages. One control stage is needed to obtain the mean value rectification and the second serves to stabilize the indicated output amplitude. In this case the amplitude modulator serves as regulating element. This control circuit is also used to adjust the output voltage with the aid of a ten-turn helical potentiometer.

The output socket II with  $Z_{out} = 75 \Omega$  is independent of the divider output and supplies the voltage available previous to the divider, thus providing a greater output voltage level for the operation of counters and similar units in addition to the attenuated output voltage.

### 2.2.1. 69 MHz Oscillator (1)

The oscillator operates as a capacitive three-point circuit. The oscillator coil consists of a single coil in the form of a metal bracket. Since the oscillator amplitude is solely determined by the emitter current operating point of the transistor, this amplitude can be adjusted with the aid of P 101. The frequency of the oscillator is held always at a constant 69 MHz by varicaps G1 101 and G1 102, which are supplied with a control voltage from the phase control (2). The buffer stages T 102 and T 103 serve to isolate the oscillator from the subsequent amplitude modulator (3), thus avoiding that a simultaneous frequency modulation can occur by interaction on the oscillator during amplitude modulation.

### 2.2.2. Phase Control (2)

This control compares the frequency of the 69 MHz oscillator to the 1 MHz standard frequency of the RWO-5 and generates the required control current for the synchronization of the 69 MHz oscillator with the control frequency.

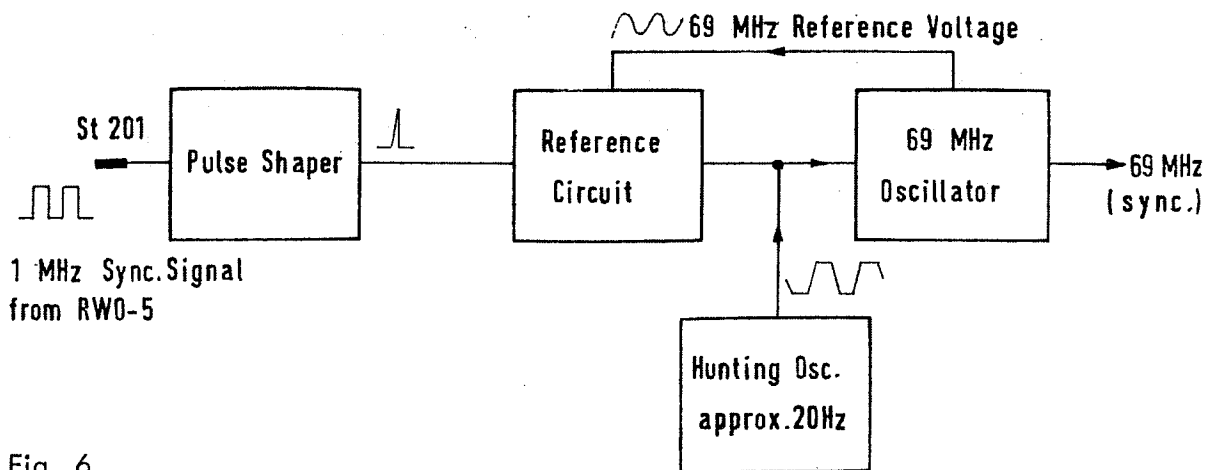


Fig. 6

The pulse shaper stage T 201, T 202 forms very narrow spikes from the 1 MHz square wave signal, whose phase position to the 69 MHz signal effects the sign and amount of control voltage.

This control voltage controls the frequency of the 69 MHz oscillator with the aid of the two varicaps diodes so that the phase position is held constant and thus synchronized. Synchronization will not occur if the 69 MHz frequency is not within the "hold range" of the phase control circuit instrument.

The hunting oscillator is, however, operative and sweeps the 69 MHz oscillator. The frequency of the 69 MHz oscillator will pass through the hold range and will thus be synchronized. The hold range has, in this manner, been increased to the value of the hunting oscillator sweep range. The hunting oscillator is so designed that oscillation ceases on loading. This is the case when synchronization is obtained because the feed point of the oscillator is made low-impedance by the effect of the control.

#### Pulse shaper:

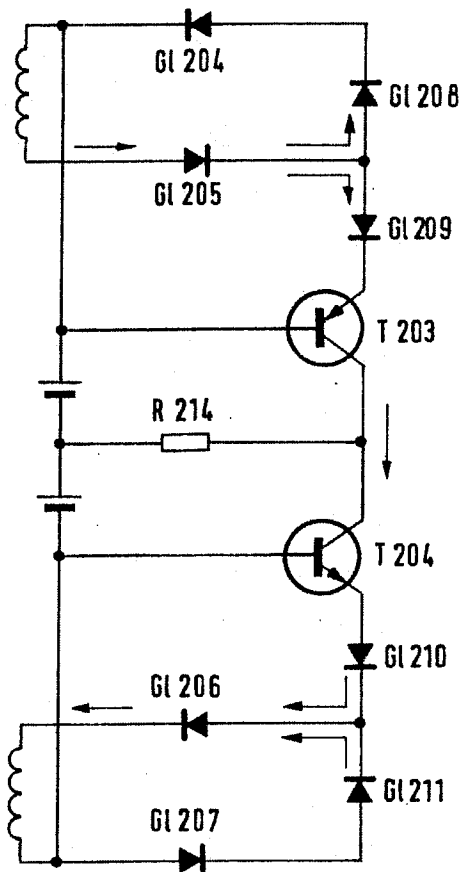
T 201 and T 202 are blocked in their quiescent state. The negative input signal is differentiated by C 201 and switches T 201. The output signal of T 201 is differentiated by C 203/R 204 and then switches T 202, whose output signal, a steep negative pulse, is differentiated by the varicap G1 202 and winding 1-2 of transformer Ü 201. A pulse with a rise time of approx. 3 ns, a width of 15 ns and a peak voltage of 7 V is formed.

#### Comparison circuit:

The comparison circuit receives the 69 MHz signal via T 206 and T 205 (approx. 0.5 V at the emitter of T 205) and the 1 MHz pulse via the secondary windings of Ü 201. T 203 and T 204 will be blocked if these two signals are not present, so that the potential adjusted at P 201 is available at the connection point of the two collectors. P 201 is normally adjusted so that the voltage between TP 202 and TP 203 is approx. zero. This condition is also maintained if only the 69 MHz signal is fed to the circuit. The transistors T 203 and T 204 are also blocked during the pauses between the 1 MHz pulses.

A current will flow via GI 205 (or GI 206) on receiving a pulse, which is branched off at the cathode (or anode).

The value flowing via GI 208/GI 209 or GI 210/GI 211 is determined by the momentary phase position of the pulse to the 69 MHz signal. The following illustration, containing only the relative information, shows the principle when the 69 MHz signal is assumed to be absent.



In a fully balanced condition half of the current flowing via GI 205 is passed through GI 209, T 203, T 204 and GI 210. Thus maintaining the common collector potential of T 203 and T 204. If, however, the current via GI 209 and through GI 210 is different, the current difference would flow via R 214 and will shift the collector voltage of one of the transistors to saturation. This could be caused for instance, by a small unbalance of the circuit.

Fig. 7

On feeding the 69 MHz signal via T 206/T 205 a current difference is generated as a measure of the phase difference. Fig. 8 illustrates three different conditions:

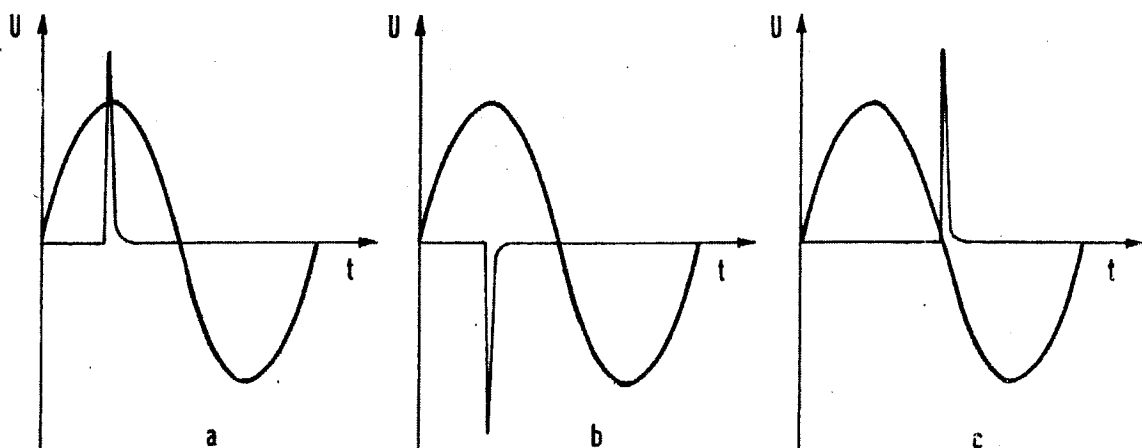


Fig. 8

Illustration a: The pulse coincides with the identical phase maximum of the 69 MHz signal, therefore doubling the current through GI 209 until GI 208 is blocked. At this moment the phase position at GI 210 is as in illustration b.

The current in GI 201 is doubled whereas the current in GI 210 is reduced.

The collector potential is shifted in a positive direction and is limited by the saturation. If the opposite phase position is present the signs are exchanged and the collector potential shifted towards minus.

Illustration c: In the phase position shown in c the average current via GI 209 and GI 210 and thus the collector potential of T 203 and T 204 are not altered. Between the phase positions illustrated in Fig. a to c, there are, naturally, all the intermediate positions. A control voltage dependent on the phase position is thus achieved. As long as the phase position is in a state of transition, a square wave voltage will be developed whose frequency is dependent on the frequency difference between the 69 MHz and the 69th harmonic of the 1 MHz synchronizing signal. The low-pass filter (R 218, C 210) ensures that the control amplification for frequencies in excess of 10 kHz is small and that the control is held stable. However, a very narrow hold range is obtained because of this, since no control voltage is generated at greater frequency differences. The collector voltage of the hunting oscillator is slowly shifted (20 Hz) back and forth to the saturation limits, thus tuning the 69 MHz oscillator. This means that the 69 MHz oscillator will be locked-in by the control voltage on reaching a frequency within the hold range and the hunting oscillator will then cease operation.

It is important that no spurious alternating voltages are present at the varicaps of the 69 MHz oscillator as these could cause a frequency modulation. The most serious interference would be that originating from the 1 MHz standard. A 1 MHz absorption circuit consisting of C 211 and L 201 ensures, in conjunction with the low-pass filter elements, an effective filtering.

Potentiometer P 201 serves to compensate for circuit unbalance, thus obtaining a hold range balanced in respect to the nominal frequency.

### 2.2.3. Amplitude Modulator ③

The amplitude modulator consists of a simple diode modulator, the principle of which is illustrated below:

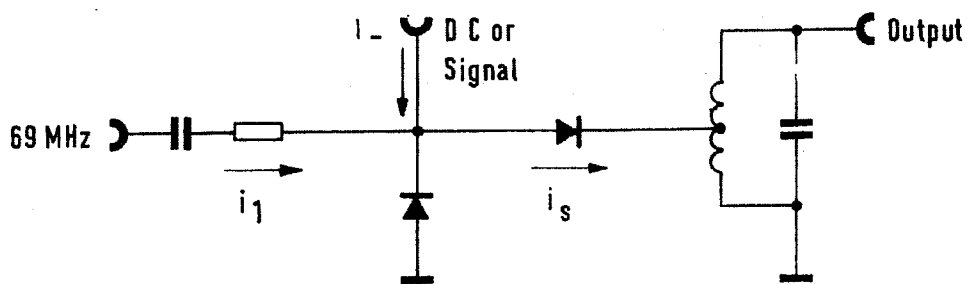
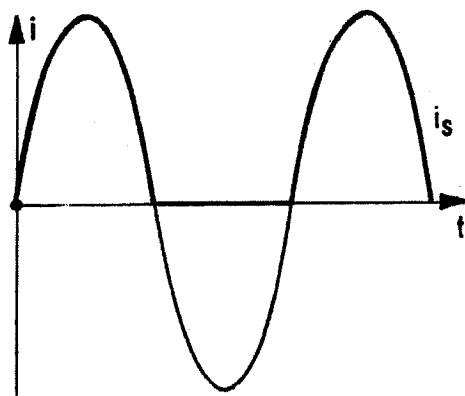


Fig. 9

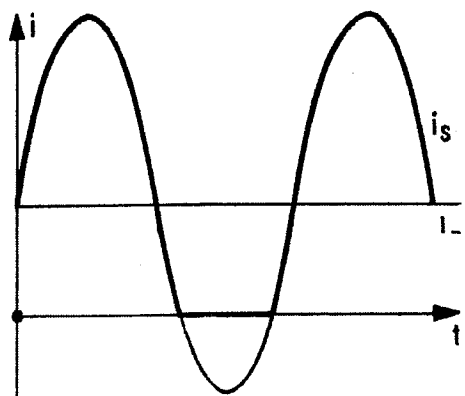
A 69 MHz sine-wave current  $i_1$  is supplied to the input. Under no current conditions at the DC input, the trace will have the illustrated form:



The current  $i_s$  in the resonant circuit is formed merely from the positive half cycle and its trace is as shown by the accentuated curve.

Fig. 10 a

If a positive DC is fed to the modulator in addition to the current shown in Fig. 10 a, it will be superimposed on to the sine-wave current  $i_s$ .



The sine-wave  $i_s$  in the resonant circuit is indicated by the accentuated curve.

Fig. 10 b

If a negative DC current is injected, then the trace will be as in Fig. 10 c

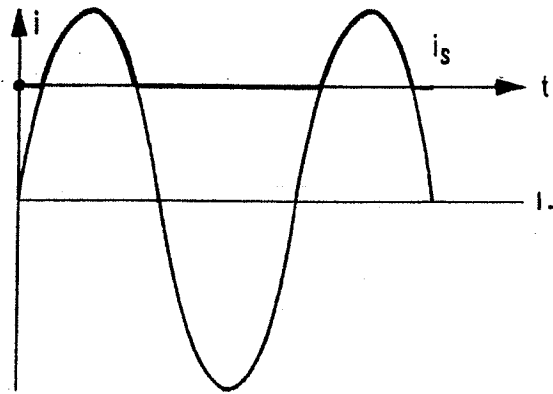


Fig. 10 c

All of the accentuated curves possess a fundamental component of the 69 MHz oscillation, in "b" more and in "c" less than in Fig. 10 a. The output voltage at the resonant circuit is proportional to this fundamental amplitude. This allowing the value of the 69 MHz to be determined by the magnitude and polarity of the injected DC. If an LF signal current is fed instead of the DC, this will have the result that the output amplitude will be varied in step with the LF, thus forming the modulation.

#### 2.2.4. 69 MHz Amplifier (4)

This amplifier essentially represents a band-pass filter for 69 MHz, which is loosely coupled via a buffer and intermediate amplifier stage. This loose coupling has the effect that with simultaneous RF amplification no dip in the filter curve is made which would be a disadvantage with regard to the modulation side bands. The filter curve is, however, so broad that the modulation side bands are passed unattenuated.

### 2.2.5. Frequency Converter (5)

The task of the converter is to form the required 10 kHz to 36 MHz frequency range of the generator portion from the variable carrier frequency (69,01 MHz to 105 MHz) of the RWO-5 and the fixed 69 MHz signal frequency generated in the PS-5.

Since the converter operates according to the ring modulator principle, the essential mixer products are  $f_{\text{carrier}} + f_{\text{signal}}$  and  $f_{\text{carrier}} - f_{\text{signal}}$ . The difference product is used in the PS-5. The sum product and all frequencies in excess of 36 MHz are suppressed in the following low-pass filter (6). The control signal is fed as carrier signal via a limiter circuit (11) to transformer Ü 501 of the converter, where it is balanced, amplified by T 501 and T 502 and passed to transformer Ü 502 of the mixer. The signal frequency is also fed balanced to the mixer. Due to the unavoidable wide tolerances, the diodes have been balanced by series resistors in the forward direction and by parallel resistors in the reverse direction. This allows the demands on the mixer in regard to carrier and signal suppression to be met.

### 2.2.6. 36 MHz Low-pass Filter (6)

This low-pass filter has the task of filtering out the difference product  $f_{\text{carrier}} - f_{\text{signal}}$  from the unwanted mixer products.

The cut-off frequency is at 36 MHz. All frequencies above 36 MHz, especially the sum products will be attenuated by more than 80 dB.

### 2.2.7. Pre-amplifier (7)

The pre-amplifier consists of 2 negative current feedback amplifier stages T 702 and T 704, which are matched to each other and to the in and output by means of buffer stages. An increase of the frequency response is made with C 705 and C 708, which can in turn be reduced by trimmer C 707. This allows the frequency response of the combination pre-amplifier and output amplifier to be adjusted.

### 2.2.8. Output Amplifier ⑧

The output amplifier has a complement of complementary transistor stages. All stages are series push-pull stages, ie. the transistors are placed in series with each other across the DC. The output stage comprises two parallel connected push-pull stages and the output impedances are formed by use of defined series resistors which operate as parallel resistors for the individual outputs.

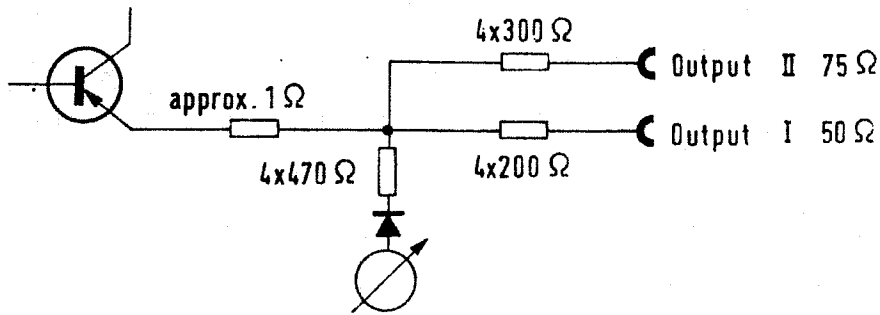


Fig. 11 Equivalent circuit diagram of one of the 4 parallel connected branches of the output stage.

The output impedance for the connection of the indicator circuit is a portion of the series resistance, which is needed for the mean value rectifier. The meter thus accurately indicates the source voltage of the output stage.

L 801 at the output of the driver stage T 801/T 802 serves to compensate for the input capacity of the following stage and to equalize the frequency response at the upper band limit. The resistor parallel to L 801 prevents an out of band response lift, which would otherwise cause an increase in the harmonic distortion at approximately 20 MHz. The DC operating point for the complete output stage is adjusted with P 801 to obtain the lowest possible harmonic distortion.

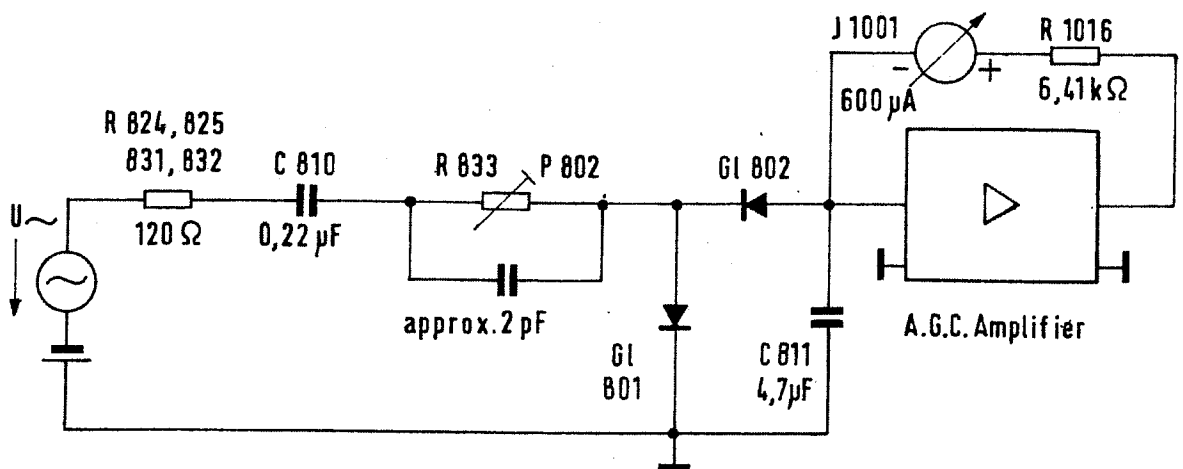


Fig. 12 Indicator circuit



If two conditions are fulfilled, a direct current proportional to the AC voltage  $U \sim$  will flow via G1 802 into capacitor C 811. These conditions are:

- 1) Capacitor C 811 must not accept any inverse voltage charge, which could cause G1 802 to conduct during the whole half wave.
- 2) The curve of the rectifier must be linearized with a sufficiently large series resistance (R 824 to R 833) in order to obtain the proportionality between current and voltage.

The first condition is fulfilled in that a control amplifier feeds C 811 with an inversed DC of equal magnitude. This compensating current is indicated on meter J 1001 and exactly corresponds to the direct current flowing via G1 802 if the control amplifier requires a negligible current and if the voltage across C 811 can be zeroed. These requirements can be made with a sufficiently high control amplification.

C 810 serves to provide a DC-free output. G1 801 prevents C 810 from being charged and additionally allows the generation of a bias current.

The threshold voltage of the rectifier will have an unfavourable effect if the AC voltage is low. Its influence on the proportionality between the alternating and direct current will then no longer be negligible. For this reason, a bias current is fed from the control amplifier to just overcome the threshold voltage. The forward voltage of the series connected diodes then appears at C 811 in the absence of an AC voltage. The condition 1) above can be condensed as meaning that no voltage variations should take place at C 811 in the presence of an AC voltage.

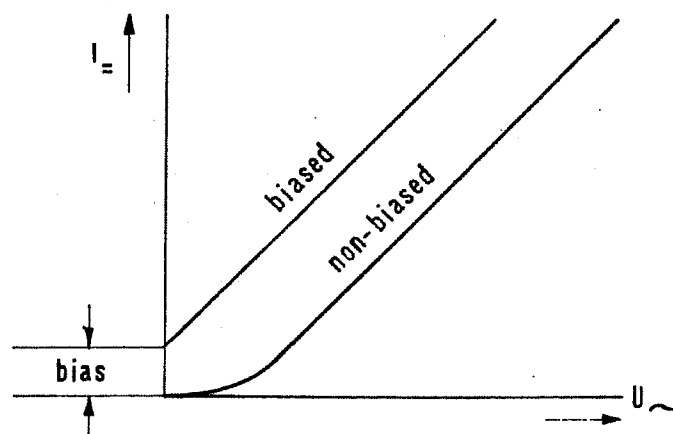


Fig. 13

Control amplifier for the indicator circuit (circuit diagram 10 )

The control amplifier comprises the transistors T 1006 and T 1005. The output voltage is available at the emitter of T 1005. Comparison is made at the input (base-emitter of T 1006) between the voltage at C 811 and a voltage obtained from P 1004. The required bias current can be selected with the aid of P 1004. Since the forward voltage at C 811 is effected by the temperature dependence of the two diodes, it is advisable to provide a comparison voltage with an identical temperature coefficient. This is achieved by the base-emitter diode path of T 1006 and by G1 1005.

#### 2.2.9. Attenuator (9)

The attenuator consists of four combinable L pads with the values 20 dB/10 dB/20 dB/20 dB or 2 Np/1 Np/2 Np/2 Np at an impedance of  $Z = 50 \Omega$ . (in circuit diagram left to right). The attenuator pads are selected by attenuator switch S 901 with the aid of reed contacts. Since high demands have to be placed on these reed contacts, gold plated contacts have been used. The energizing coils of the reed contacts are partly switched via diodes and due to this it was possible to use a uncomplicated attenuator switch. The resistors R 909 to R 913 are simultaneously switched with output impedance switch S 902 and are used to switch the meter sensitivity in the "power level" position.

#### 2.2.10. Amplitude Control (10)

Any variation of the mean value of the AC output voltage will cause a proportional change in the DC voltage at the emitter of T 1005 as well as a variation of its collector voltage. The amplitude control makes, with the aid of the difference amplifier T 1003/T 1004, a comparison between the collector voltage of T 1005 and the DC voltage selected at P 1002, and equalizes same. If the voltage selected with P 1002 is varied, the output AC voltage will also be altered. The control circuitry needed for this consists of the difference amplifier, amplitude modulator (3), 69 MHz amplifier (4) converter (5), 36 MHz low-pass filter (6), pre-amplifier (7) and output stage (8) with rectifier circuit.

A rise in the collector current of T 1003 will cause an increase of the AC output voltage, whereas a current decrease will cause a fall-off of same.

With the aid of P 1001 the emitter voltage of T 1001 is adjusted, thus controlling the collector current of T 1002 as well as the voltage drop across the level control P 1002. The voltage drop across P 1002 is adjusted so that at the fully clockwise position of P 1002 the f.s.d. of the meter is accomplished (The emitter voltage of T 1001 is adjusted to 4 V on alignment). The function of P 1002 is still maintained even if the control circuit is broken at socket Bu 1001 because the collector current of T 1003 and thus the AC output voltage are effected by this control. A modulation voltage at socket Bu 1001 would also vary the collector current of T 1003 and thus cause an amplitude modulation of the output voltage. If one ensures, with the aid of diode G1 1003, that the collector circuit of T 1002 has the same voltage threshold as the input circuit with T 1003, then the collector current of T 1003 will vary proportionally to the position of potentiometer P 1002 or to the modulation voltage at socket Bu 1001. G1 1004, in the other portion of the difference amplifier (T 1005), is necessary for reasons of balance and to achieve a low temperature coefficient.

The operating point of the amplitude modulator can be varied with potentiometer P 1003. P 1003 must be so adjusted that T 1004 passes current at all times to ensure that the control amplifier is not overloaded within the whole output voltage range. A second condition of adjustment is the correct position of the control range of P 1002 if the control circuit is non-operational (amplitude modulation).

If the amplitude control stage is inoperative and a resistor is placed in parallel with the link consisting of J 1001, R 1016 and P 1005, the emitter voltage will sink until the current through the meter has been reduced by the value of the addition current flowing through the parallel resistor.

To summarize: The emitter voltage is so reduced that the compensating current originally flowing to C 811 continues to maintain its value. This naturally has no effect on the output level of the output stage. If, however, the amplitude control stage is operative, the output level will continue to increase until the original emitter voltage of T 1005 available before switching of the reference level is reached and thus allowing the original meter current to flow once again.

If the correct parallel resistor has been chosen, the result will be a constant level reading with the corresponding correct output level. Since the level range indication is simultaneously increased by +10 dB (+1 Np), only small corrections of the output level are necessary on switching from voltage to power level calibration. This has a favourable effect on the linearity of the indicator circuitry on switching its sensitivity.

### 2.2.11. Limiter (11)

The two series connected difference amplifier stages operate as a limiter and have the task of limiting the control signal from the RWO-5. This allows the mixer to be fed with a constant carrier signal independent of the amplitude of the control signal.

### 2.2.12. Voltage Filtering (12)

The voltage filtering suppresses high-frequency voltages which are induced into the operating voltage lines of the generator portion.

### 2.2.13. Hunting Oscillator (13)

The hunting oscillator should commence operation as soon as the 69 MHz oscillator is not in its locked condition. The 69 MHz oscillator then supplies a 20 Hz square wave signal which shifts the 69 MHz oscillator into the hold range. The following simplified circuit diagram illustrates the principle of this circuit.

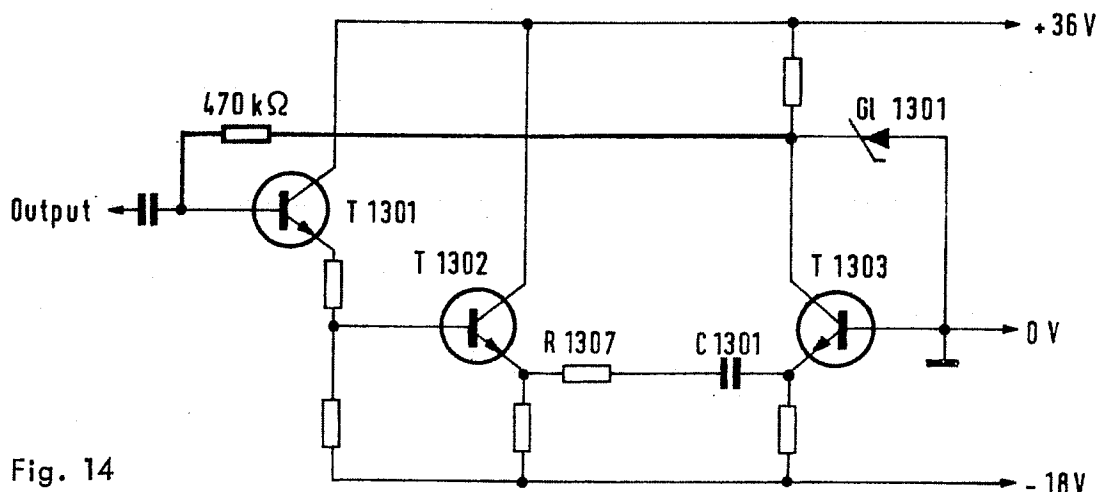


Fig. 14

The line shows the feed-back path connecting the output transistor T 1303 with the input of the circuit. Due to the load resistor at the output, the coupling factor  $k$  of the feed-back will be effected by the  $470\text{ k}\Omega$  series resistor and oscillation will cease as soon as  $k \times V \leq 1$ . Due to the high impedance of the feed-back path, this is the case at approximately  $70\text{ k}\Omega$ .

If the  $69\text{ MHz}$  oscillator is not synchronized, the load impedance will amount to approx.  $180\text{ k}\Omega$ ; whereas on synchronization this value will be far lower than  $70\text{ k}\Omega$ . The hunting oscillator will therefore be switched in and out, since its amplification  $V$  is held sufficiently constant by the corresponding circuitry. The oscillator frequency is determined by R 1307 and C 1303. GI 1301 limits the output amplitude to a defined value.

### 2.3. Receiver Portion PM-5

#### Operation of the complete unit

The receiver portion PM-5 is always part of a combination. It is either combined with the Oscillator RWO-5 to form a complete Receiver PMO-5 or in conjunction with the Oscillator RWO-5 and Generator PS-5 to constitute the Level Measuring Set PSM-5. The oscillator Portion supplies the control signal, a synchronizing signal and the stabilized operating voltages. The receiver allows selective measurements to be made in the range from 10 kHz to 36 MHz. The bandwidth is switchable between 500 Hz and 3.5 kHz. The input impedance of the receiver is either high-impedance or any of six different impedances between 50  $\Omega$  and 150  $\Omega$ . The measured voltage can be indicated as either a linear or logarithmic reading. In addition to this the measurement magnitude can be indicated as a voltage or power level for the available input impedances both with and without terminating resistor. The same meter scale is valid both for the voltage and power level, which has been achieved by correspondingly switching the sensitivity. The dB model also possesses an additional volt calibration.

The wide-band portion of the PM-5 comprises the pre-attenuator for the selection of the sensitivity, buffer and amplifier stages. The first intermediate frequency of 69 MHz is generated in mixer I as the difference of the RWO-5 control frequency (69.01 MHz to 105 MHz) and the receive frequency (10 kHz to 36 MHz). Since the IF is "up-converted", the low-pass filter preceding the mixer ensures a high image rejection. The final intermediate frequency of 30 kHz is obtained after three further frequency conversions. It is at this final IF where the actual selection of the receiver is carried out. The conversion frequencies generated in the RWO-5 are phase locked to another and are synchronized to the crystal spectrum of the RWO-5, thus assuring that the frequency stability of the PM-5 is identical to that of the RWO-5. The signal is then passed via the final amplifier and output-attenuator to the output sockets and to the indicator circuitry. The receiver can be switched to the low-distortion or low-noise mode for the two extreme modes as sweep-receiver or distortion analyzer. Since no attenuation takes place in the most sensitive position of -110 dB (-12 Np), there will be no difference between the low-noise and low distortion measurement. If the sensitivity is to be reduced, then this is made by firstly operating the output-attenuator for low-noise mode or the pre-attenuator in the low-distortion mode.

The receiver possesses output sockets for the 30 kHz IF signal, for a DC output and for an output for headphones.

Voice channels in either the normal or inverted position are placed into their original position by mixing with a built-in frequency conversion oscillator. Double sideband signals are demodulated in the normal manner. The output sockets are floating and fully independent of another. They can be loaded up to short circuit. Sockets are also provided for the accessory Display Unit SGE-10, Scale Expander AZD-1 and Probe TK-8.

### 2.3.1. Input Circuit (1)

The input circuit consists of the terminating resistors, the pre-attenuator and a cathode follower stage comprising the only tube of the Level Measuring Set PSM-5.

The terminating resistors can be switched to the input by the reed contacts of relays 101 to 106. The compensation windings of the adjacent relays are simultaneously switched thus eliminating the effects of the magnetic flux, which could have a disturbing effect on the neighbouring relays, due to the close-spaced assembly. The addition of a low-pass filter has enabled the input capacitance of the unit to be compensated for the impedance values 75  $\Omega$  and 150  $\Omega$ .

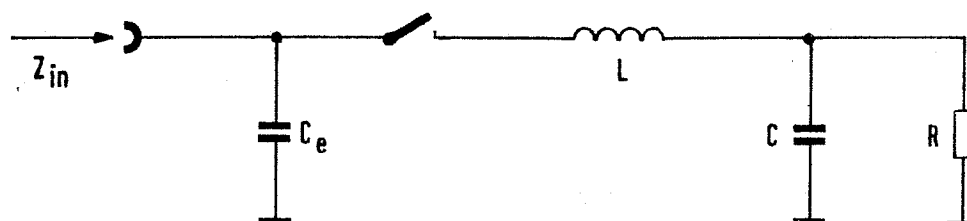


Fig. 15 Terminating resistor

The other Z values are formed by switching a further resistor in parallel.

A 40 dB (4 Np) capacitive divider consisting of C 107 parallel to C 108 and C 110, precedes the cathode follower stage. The input capacity is increased to approximately 30 pF by C 104 parallel to C 106. This value of 30 pF corresponds to the value which would appear if the grid of the tube is directly connected. The capacitive impedance will change at frequencies exceeding 15 MHz and will be also dependent on the setting of the pre-attenuator.

The triode connected valve E 810 F is a low-noise type possessing a mutual conductance

of 50 mA/V. A control circuit comprising transistor T 1901 maintains the valve current at a constant 40 mA.

This control circuit is not part of the input unit but is contained within the printed circuit board "main distributor 243-B". Any low-frequency oscillation of the control stage is neutralized by C 1907<sup>1)</sup>. The combination of R 110 and C 116 in the anode lead compensate the negative real component of the cathode-follower input impedance in the case of capacitive loading. The grid neutralizing resistor can therefore be kept at a low value, so that no additional noise is added to the inherent noise of the circuit. The chosen control circuit allows a high grid leak resistance (R 107) of 10 M $\Omega$ , which essentially represents the input impedance of the valve stage at low frequencies. The cathode resistor with a total of 300  $\Omega$  is dimensioned as a resistive voltage divider of 0, 10, 20 or 30 dB (0, 1, 2 or 3 Np). The pre-attenuator and the capacitive divider preceding the cathode follower allow the selection of attenuation levels of 0 and 10 to 70 dB (0 and 1 to 7 Np) in steps of 10 dB (1 Np). The coupling to the pre-amplifier is made via two isolating capacitors, which ensure that the relatively slow-switching gas-pressure relays do not cause a momentary short-circuit of the cathode resistor. The taps for 0 dB and 20 dB (0 Np and 2 Np) attenuation are made with series resistors to increase the source impedance to the value of the 10 dB (1 Np) position, thus ensuring that the preamplifier is provided with a constant source impedance. The 30 dB (3 Np) attenuator is calculated for the same source impedance (69  $\Omega$  for the dB model or 72  $\Omega$  for the Np model).

The frequency response of the various attenuator positions are aligned with trimmer capacitors C 117, C 119 and C 120. Chokes L 103, L 104 and filter resistor R 120 suppress the injection of spurious signals from the RWO-5 via the power supply lines. The calibration signal is fed via relay 112 direct to the grid of the tube. The input isolated by relays 107 and 109.

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<sup>1)</sup> Series A to C: C 129



### 2.3.2. Calibration Oscillator 100 kHz (2)

The calibration signal is generated in a low-distortion transistor oscillator stage. The oscillator operates according to the following principle:

A stabilized voltage is momentarily fed to a high Q resonant circuit on passing through its maximum amplitude and is then allowed to oscillate free during the rest of the period. Transistor T 201 operates as a switch and obtains its control voltage via a feed-back winding of the resonant circuit transformer. The phase, the current flow angle and the magnitude of the commencement pulse are determined by the values of the capacitors and resistors in the base circuit.

The sine-wave calibration signal is tapped off, at a relatively low impedance, from a further winding of the resonant circuit transformer. A voltage divider, possessing a negative temperature coefficient (thermistor R 206), compensates for the temperature coefficient of the zener diode G1 201, which directly controls the amplitude of the oscillation by means of its zener voltage. The calibration signal is available at two output sockets, one at a level of -20 dB and a second at -3 Np. The exact value is adjusted with potentiometer P 201. The oscillator frequency of 100 kHz is aligned at transformer Ū 201. In the production series A to E, the calibration oscillator is fed with its operating voltage only if one of the calibration push-buttons are depressed. Commencing with series F, the calibration oscillator is, in addition, switched on with a slight delay via the base of T 201.

### 2.3.3. Pre-amplifier 10 kHz to 36 MHz (3)

The pre-amplifier consists of an amplifier stage (T 302) preceded by a common-collector stage (T 301) and succeeded by a difference amplifier (T 303, T 304). The emitter resistor of the common-collector stage is at the emitter of T 302 and is thus also effected by the fluctuations of the input voltage. The emitter resistor is thus effectively larger and the input impedance of the pre-amplifier of higher impedance. Due to the alternating switching of the two transistors T 303 and T 304, the difference amplifier has the effect of suppressing the second harmonic. The voltage drop across R 307 is fed back from the collector of T 303 to the emitter of the amplifier stage. This feedback is frequency dependent and allows a compensation of the basic frequency response inductance by means of L 301 - which is dependent on the input unit. The transistor T 304 feeds the 36 MHz low-pass filter at an output impedance of 150 Ω.

The operating point of T 301 is determined by alignment R 314 so that the difference amplifier possesses a balanced current division in the absence of an input signal. Diode G1 301 protects the base-emitter path of T 301 against break-down, should a high AC voltage or peak voltage reach the pre-amplifier from the input. The upper frequency limit is mainly determined by the frequency limit of the current amplification of the transistors.

#### 2.3.4. 36 MHz Low-pass Filter (4)

This low-pass filter suppresses all frequencies in excess of the receive frequency band. It is designed for lowest ripple within the pass-band. The three poles ensure a high degree of suppression of the first intermediate frequency of 69 MHz and the image frequency (receive frequency plus 138 MHz).

#### 2.3.5. Mixer I and 69 MHz Amplifier (5)

The receive frequency of 10 kHz to 36 MHz is converted in mixer I, with the aid of the RWO-5 control signal ( $f = 69,01$  to  $105$  MHz), to the first IF of 69 MHz. The IF level is then increased in the subsequent amplifier.

The mixer represents a  $150 \Omega$  terminating impedance for the 36 MHz low-pass filter. The two diode pairs switch the input signal  $\omega$  to the output during a half-cycle of the control frequency  $\Omega$ . The main mixing products are thus the sum and the difference resulting from the control and receive frequency.

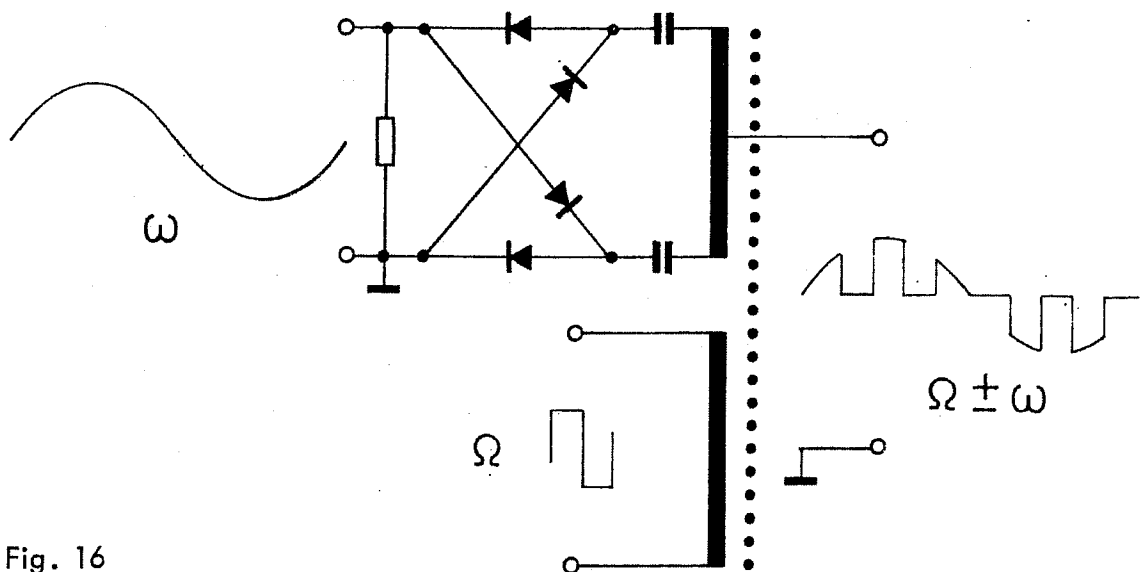


Fig. 16

The difference between both frequencies results in the first intermediate frequency of 69 MHz. Since the carrier level is substantially higher than the signal level, it could reach the output and overload the succeeding stages, if the mixer was in an unbalanced condition. For this reason it is possible to capacitively balance the carrier transformer Ü 501 with the trimmers C 508 and C 510. Differing diode thresholds can be compensated with the aid of a bias current via R 512 and thus allowing the mixer to be resistively balanced. The current is adjusted with the potentiometer P 2001 mounted on the "Distributor 2" printed circuit board.

The potentiometer is also accessible from the rear of the cabinet. R 516 and R 517 represent linearizing resistors for the characteristic curves of the diodes. The 69 MHz amplifier possesses a relatively high input impedance so that the mixer is not loaded. The feed-back resistor R 508 of amplifier stage T 501 is placed across a resonant circuit to ground offering a certain selectivity. The common-base stage consisting of transistor T 501 with its load resistor R 505 operates into the 69 MHz band-pass and suppresses any retroaction of the latter. The operating voltage of the common-base stage is fed at low impedance via transistor T 2001.

#### 2.3.6. Limiter 69.01 to 105 MHz (6)

The RWO-5 control signal is converted into a square-wave and limited within the limiter to obtain the lowest possible dependence of the dynamic slope of mixer I on the carrier voltage. The control signal is fed via a low-pass filter, which operates as a image suppression filter, to the limiter stage consisting of T 601 and T 602 <sup>1)</sup>. The amplitude of the drive is sufficiently large to fully actuate this stage. The current determined by the common emitter resistor R 605 flows thus solely through one transistor during the first half cycle, whereas in the second half cycle only via the other transistor. The inductance L 601 equalizes the capacitance at the collector of T 602. The second limiter stage comprising transistors T 603 and T 604 generates a sufficiently large square-wave with a high degree of steepness. The mixer diodes G1 501 to G1 504 are thus fully opened via the carrier transformer Ü 501.

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<sup>1)</sup> The low-pass filter was replaced by an isolating transformer in series A to C.

A control voltage of  $U_{r.m.s.} \approx 0.5$  to a maximum of 2 V can be fed to socket "CONTROL INPUT PM-5" of RWO-5, if it is desired to tune the unit with an external oscillator. The receive frequency will then be at 69 MHz below the control frequency.

### 2.3.7. 69 MHz Band-pass Filter (7)

This band-pass filter consists of three basic  $\pi$  links, which are inductive or capacitively coupled to another. Its task is to filter out the first intermediate frequency of 69 MHz from the mixing products of mixer I, whereby the suppression up to 500 MHz must be sufficiently large to ensure a high level of interference rejection. An attenuation peak at 54.12 MHz (C 711) suppresses the image frequencies of the succeeding mixer, which are  $2 \times 7.44$  MHz below the first intermediate frequency. The input of the filter is unbalanced; the output balanced for the driving the transistor mixer II.

### 2.3.8. Mixer II and Buffer Stage (8)

Common-collector stages comprising transistors T 801 and T 802 are provided before mixer II to suppress retroaction from the mixer to the 69 MHz band-pass filter. The base resistors R 801 and R 802 represent the terminating resistors for the band-pass filter. The neutralizing resistors R 804 and R 807 suppress any instability of the two common-collector stages. Mixer II converts, with the aid of a 61.56 MHz carrier oscillator, the first intermediate frequency of 69 MHz to the second intermediate frequency of 7.44 MHz.

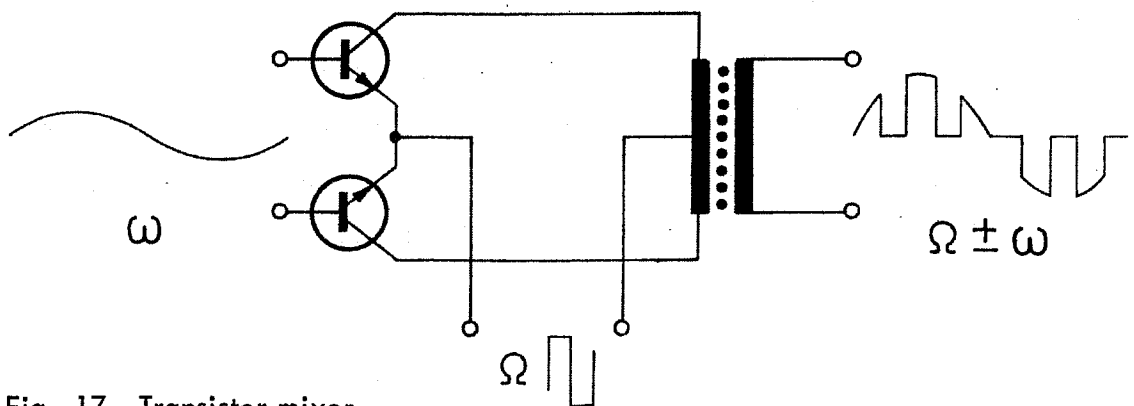


Fig. 17 Transistor mixer

The mixer transistors T 803 and T 804 are switched by the square-wave carrier current. This current is generated in a transistor limiter driven by the 61.56 MHz oscillator. The 69 MHz intermediate frequency only appears at the output of the mixer during the conducting phase of the transistors, which means that the sum and the difference of the first intermediate frequency and carrier frequency are generated. The succeeding band-pass network filters out the difference frequency of 7.44 MHz. The carrier current then flows back through the band-pass transformer and cancels itself out. An intermediate frequency amplifier is not necessary, because the mixer transistors produce a certain degree of amplification during the conducting phase. The task of resistor R 817, at the carrier input of the mixer, is to ground the emitter potential of T 803 and T 804 during the current-free half-cycle. This ensures that the switching point is rapidly reached. Both this and the negative feed-back via the emitter resistors contribute to a low-degree of harmonic distortion.

#### 2.3.9. 7.44 MHz Band-pass Filter (10)

The band-pass filter comprises one main and two adjacent links. The main link is divided into two parts. The adjacent links form an attenuation peak at 6.25 MHz, which is the image frequency of the succeeding mixer, appearing at  $2 \times 600 \text{ kHz}$  below the second intermediate frequency. The two band-pass transformers Ü 1101 and Ü 1102 ensure a balanced in and output.

#### 2.3.10. Mixer III (11)

This mixer is built up and operates in the same manner as mixer II and converts, with the aid of a carrier frequency of 6.85 MHz, the second intermediate frequency to the third IF of 600 kHz, which is filtered out in the succeeding 600 kHz band-pass filter.

#### 2.3.11. 600 kHz Band-pass Filter (11) (12)

This band-pass filter is assembled in the same manner as the 7.44 MHz band-pass filter and its attenuation peak is at the image frequency of the following mixer, which is at  $600 \text{ kHz} - 2 \times 30 \text{ kHz} = 540 \text{ kHz}$ .

attenuator output is shorted by relays 1501 and 1502, thus avoiding a voltage-overload of T 1502 at high levels.

In the linear mode, T 1502 also operates into the output-attenuator. In conjunction with matching capacitors, the two transformers  $\dot{U}$  1501 and  $\dot{U}$  1502 form parallel circuits resonant at 30 kHz. The damping resistor across  $\dot{U}$  1501 and thus the sensitivity of the whole unit is controlled with potentiometer P 1501 "CAL lin".

The emitter resistor of T 1502 is varied according to the Z value by switch S 101 in the "POWER LEVEL" position. The feed-back is dimensioned so that the same meter scale can be used. The attenuation scale will additionally be shifted by 10 dB on when switching to the power level mode. The unit is automatically switched to voltage level on depressing the calibrating button. The output-attenuator allows the attenuation of 0 dB, 10 dB to 70 dB (0 Np, 1 Np to 7 Np) and with the aid of the transformer divider stages 10 dB, 20 dB and 40 dB (1 Np, 2 Np and 4 Np). Diodes GI 1501 to GI 1506 allow the various attenuator combinations to be switched with a minimum of switch contacts.

#### 2:3.15. 30 kHz Amplifiers 2 and 3 (16)

These two amplifiers possess essentially the same circuit as amplifier 1. However since the amplification is greater, it was necessary to place an additional common-collector stage in the circuit in order to avoid a too greater capacitive loading of the input common-collector stage. The current amplification of this transistor serves to increase the input impedance. This is necessary to ensure that the output-attenuator is not loaded in the linear mode. The high output impedance of the output transistor T 1608 is required because the output circuitry is to be fed with impressed current. The neutralizing resistors R 1621 and R 1622 and the common-collector stages T 1604 and T 1607 avoid a negative input impedance of the amplifier. A parallel low Q resonant circuit is to be found between the two amplifiers to short-out all frequencies not adjacent to 30 kHz. The antiparallel diode at the L 1601 tap limits the drive to the final amplifier and ensures that the meter pointer remains at f.s.d. when overdriving.

Transformer  $\dot{U}$  2601 is shorted by relay rel 1604 in the linear mode to ensure that no 30 kHz signal voltage can be induced into the high-impedance input.

on whether the sideband is in its normal or inverted position. In the "NARROW BAND" position the sideband signal is converted to approx. 1 kHz by a carrier frequency of 31 kHz or 29 kHz. The output voltage corresponding to f.s.d. of the meter will amount to approx. 500 mV for a converted single sideband signal and approx. 150 mV for a demodulated double sideband signal ( $m = 30\%$ ).

### 2.3.17. Logarithmic circuit (16)

Silicon diodes exhibit a strictly logarithmic relationship between current and voltage in the low forward voltage range. This effect is used to obtain the desired logarithmic values at diodes G1 1601 and G1 1602. The differential diode resistance and the corresponding voltage drop increase in a logarithmic manner with a linear increase of current.

Relays P1601 to P1604 are energized in the logarithmic mode. The 30 kHz amplifier 15 feeds the current  $i$  via the diodes, which are antiparallel-switched in regard to AC voltages. The AC voltage drop  $u$  across the diodes is amplified in stages 2 and 3 and is subsequently rectified. The following DC amplifier supplies a bias voltage  $U$  to the diodes, which is proportional to the output magnitude. The operating point on the characteristic curve of the diodes is so chosen that the voltage drop across the differential resistor will be  $U \sim \log i$ .

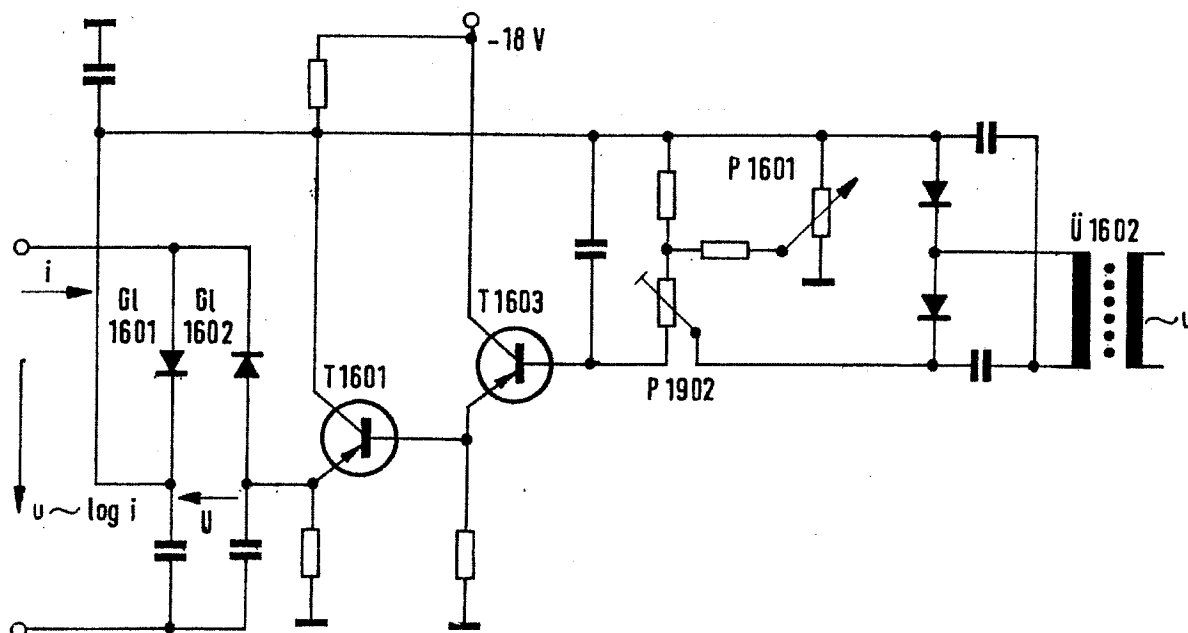


Fig. 18 DC amplifier and logarithmic diodes

### 2.3.19. Switch Assembly and Level Switch (18)

The switch contacts are mainly used to switch the heater circuits of the gas-pressure relays. The PM-5 is, with the exception of the terminating resistor switching, fully equipped with this type of relays. The heater voltage of 3 V must remain within certain limits to ensure short switching periods (less than 1 sec.).

The number of energized relays depends on the selected mode and sensitivity of the level meter. Due to the max. current drain of 0.8 A a marked voltage drop would be caused at the various filter chokes, switch contacts, switches, etc. Resistors R 1802 to R 1805 are used as loads simulating the unenergized relays.

Various relay heater circuits are intercoupled with diodes (or transistors switched as diodes) in order to reduce the number of switch contacts. The other diodes placed before relays in series F and subsequent series are solely provided that the heater voltage is reduced by the value of the diode threshold voltage, thus ensuring that all relays obtain the same energizing voltage.

### 2.3.20. Main Distributor (19)

The main distributor possesses filter circuits for all of the voltages connections including the ground lead, originating in the RWO-5. This ensures that no interfering voltages can be induced into the PM-5 via the RWO-5 power supply cables. Choke L 1907 has a winding for each voltage so that the sum of the flowing currents is zero and that the core is not pre-magnetized or saturated.

The ground of the PM-5 is RF isolated from that of the RWO-5. The series circuit consisting of R 1915 and C 1908 suppresses any resonance of the transfer impedance between the two grounds.

### 2.3.21. Distributor 2 (20)

All sub-units of the tiltable chassis are supplied from this distributor with voltages obtained from the main distributor. This is carried out partly by means of a cable harness. The common-base circuit T 2001 represents a low impedance voltage source for the 69 MHz amplifier. The carrier balance of mixer I is aligned with the trimmer potentiometer P 2001.



Capacitor C 1303, which has been previously negatively charged via R 1302, is placed at ground potential during the negative pulse. A positive synchronizing pulse is fed to the emitter of the phase comparison stage. The pulse duration should not be longer than a half-cycle of the 570 kHz signal. If it were longer, a portion of the pulse would appear at the positive half-cycle. This would mean that the decoupling transistor would never quite be blocked and one could thus not take advantage of the voltage deviation at the collector. The current pulses through T 1301 charge capacitor C 1302, whose voltage is used to control the 570 kHz crystal oscillator. The control voltage is superimposed with a 10 kHz sawtooth voltage caused by the pulse type charge and the discharge via the load resistor R 1303. This sawtooth must be filtered out in a low-pass filter to avoid an undesired modulation of the 570 kHz signal. The common-collector stage comprising T 1304 also suppresses such a modulation and feeds the 570 kHz signal to the base of T 1303. The amplitude of this voltage has been so chosen that it is in its self not sufficient to unblock transistor T 1303.

#### 2.3.22.2. 570 kHz Crystal Oscillator (12)

This oscillator generates carrier frequency IV which is used to control mixer IV. The frequency is locked to the 57th harmonic of the 10 kHz standard frequency of the RWO-5.

Since quartz crystal control is used it is only necessary to have a small hold range and the PM-5 could operate without a synchronizing signal, however, with a somewhat lower frequency accuracy.

The oscillator transistor T 1210 together with T 1209 form an emitter coupled limiter stage. T 1209 supplies the square-wave control current to mixer IV. A parallel resonant circuit comprising C 1226 and  $\dot{U}$  1203 is provided in the collector circuit of T 1210 to filter out the fundamental signal. This voltage is fed-back via various components from the secondary of the transformer to the base of T 1210. The magnitude is sufficient to fully drive the limiter.

The control magnitude is fed via a common emitter stage (T 1213) and is used as bias current for diode G1 1204. The diode can thus be shifted from the blocked into the conducting range. The effect on the oscillator frequency is explained in the following limit conditions:

As is known,  $\omega = d\varphi / dt$  is valid. The phase shift in the low-pass filter and output resistor of the decoupling transistor T 1202 must remain below  $90^\circ$ , otherwise an oscillation could occur within the hold range of the circuit if the ring amplification was sufficiently high. The charge capacitor C 1211 is placed across a resistor at the collector of the decoupling transistor T 1202 to neutralize this effect, which causes a retro-phase shift above the cut-off frequency of the RC link. It is true that capacitor C 1210 of the low-pass filter will be more quickly charged, but the charge is passed on to C 1211 from which the control voltage is taken. The second half of the low-pass filter is to be found in the subsequent sub-unit. Capacitors C 1130 and C 1206 complement the coil capacitance of the two low-pass filter coils so that attenuation peaks appear at 570 kHz and 1.71 MHz. These two frequencies are strongly contained within the synchronizing signal.

#### 2.3.22.4. 6.84 MHz Oscillator (11)

This LC oscillator generates carrier frequency III for the control of mixer III and its frequency is locked to the 12th harmonic of the 570 kHz crystal frequency.

The oscillator transistor T 1105 forms, in conjunction with T 1102, an emitter-coupled limiter stage. A parallel resonant circuit comprising transformer  $\ddot{U}$  1103 is to be found in the collector stage of T 1105. The secondary of this transformer is used to feed-back a portion of the fundamental voltage to the base. The resonant circuit capacity and thus the oscillating frequency is controlled by a control voltage at the varicap diode G1 1101.

The third harmonic of the 6.84 MHz signal is filtered out at transformer  $\ddot{U}$  1104 and is fed via the 20.52 MHz band-pass filter to the following frequency locking device. The square-wave current of transistor T 1102 flows via the common-base circuit T 1101 to mixer III and switches the transistors at the carrier frequency. The common-base circuit ensures that no interfering frequencies originating from the mixer can modulate side frequencies on to the carrier.

#### 2.3.22.5. Frequency locking device for 20.52/61.56 MHz (10)

This stage comprises essentially the same phase comparison circuit as the frequency locking device for 10/570 kHz. The generated control voltage is used to lock the 61.56 MHz oscillator to the ninth harmonic of the 6.84 MHz oscillator.

## 2.4. Scale Expander AZD-1

### Application:

The scale expander allows portions of the meter scale ( $\pm 1$  dB or  $\pm 0.1$  Np) to be spread over a length of approximately 100 mm. The attenuation range of the AZD-1 overlaps the 10 dB or 1 Np steps of the PM-5 level selector, i.e. the range of indication for each position of the level selector amounts to -10 to +2 dB (-1 to +0.2 Np). Thus obtaining an overlap of 2 dB (0.2 Np)

### Circuit:

The circuit consists of the differential amplifier T 1 and T 2, the two current amplification stages T 4 and T 5, the meter J 1 and the switchable attenuator R 9 to R 20.

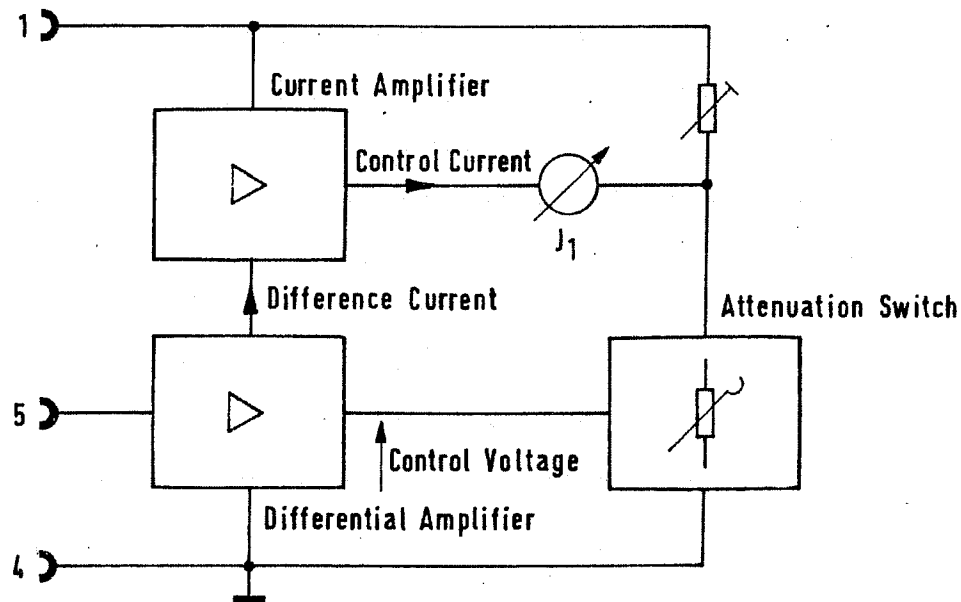


Fig. 19

The differential amplifier, current amplifier, meter and attenuator form a control circuit. A control current corresponding to the input voltage is used to bring the base voltage of T 2 to that at the base of T 1. This control is proportional to the input voltage and is indicated on meter J 1, which is calibrated in dB or Np. A complete balance of the differential amplifier is indispensable for a good linearity of the indication. The current must therefore be distributed equally between both transistors on driving the differential amplifier.

## 2.5. Probe TK-8

The Probe TK-8 is used during measuring tasks, when it is desired to make low-capacitance measurements or if it is necessary to avoid mismatch errors.

The circuit consists of two main parts:

Part 1 is formed by the two isolating stages, which are installed in the probe tip.

Part 2 comprises an additional isolating stage and controls for the matching, frequency response and the residual attenuation adjustments. This part is to be found in the cylinder which is screwed directly on the PM-5 input. The two parts are interconnected by means of a 75  $\Omega$  cable.

The emitter resistor of T 1 is connected to the emitter of T 2. This ensures that equal-phase signals appear at both ends of R 4 and R 3, which, due to the common-collector stage amplification of  $V \approx 1$ , possess virtually the same amplitude. The result is that only very small alternating currents will flow through the two resistors and that the input impedance of T 1 will appear high-impedance.

R 1 is parallel to this input impedance and determines the magnitude of same in the lower frequency range.

The input capacitance remains at a constant 7 pF over the whole frequency range.

A protection diode GI 1 is placed parallel to R 3, which ensures that the base-emitter paths of transistors T 1 and T 2 are not overloaded by large negative voltages. C 1 is provided to separate the measuring voltage from a superimposed DC voltage.

The diode GI 1 is blocked during normal operation, by the threshold voltages of the two transistors. This diode has no effect on the frequency response or input capacitance because equal-phase signals appear at both ends.

Part 2 of the circuit is matched to the 75  $\Omega$  cable by means of P 1. This adjustment is made together with the frequency response alignment to obtain the most linear curve over the whole frequency range. Trimmer capacitor C 2 is used to align the frequency response of the probe to the Receiver PM-5.

Transistor 3 also operates as a common-collector stage and feeds the signal via C 4 to the receiver input. The input impedance of the PM-5 should be switched to position  $R_e = \omega \frac{1}{C_e}$  during probe operation. The probe then possesses an attenuation of 2 dB (0.2 Np), which can be adjusted to its nominal value with the aid of P 2.

Diode GI 2 protects transistor T 3 against excess voltages.

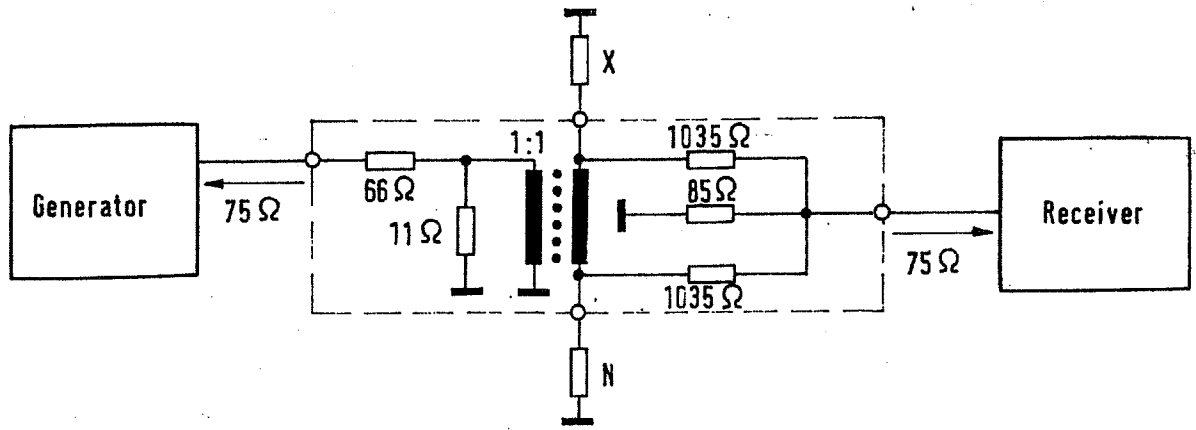


Fig. 21 Dimensioning of the bridge

The relationship of the resistors of the bridge have been calculated in such a manner, that a direct proportionality between  $U_L$  and  $r$  exists at  $N = 75 \Omega$ . An error occurs at all other resistive values, which is directly proportional to the magnitude of the reflection coefficient.

An error of  $r = 0.05$  is obtained at  $N = 50 \Omega$  and  $r = 0.086$  at  $N = 150 \Omega$ .

### 2.6.2. Circuit

The input attenuator provides a balanced output to avoid an unbalanced field distribution. Frequency response errors of the bridge, generator or receiver can be compensated for by capacitive alignment at C 101.

The auxiliary winding 5-6 is inversely poled to the primary winding 7-8, thus obtaining a symmetrical stray capacity distribution on the decoupling winding 3-4.

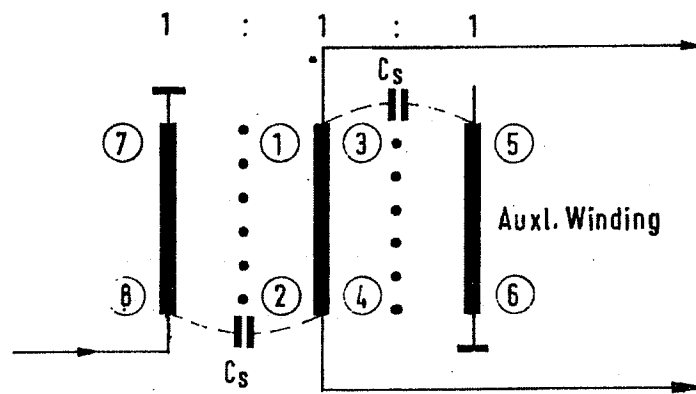


Fig. 22 Transformer Ü 101

The variable differential capacitor C 102 and potentiometer P 101 enable the bridge to be capacitively aligned. Resistors R 105 and R 106 neutralize a possible resonance between C 102, P 101 and the attached resistors or the test object.

The high impedance connection to C 103 has the effect of decreasing the capacitive load of the transformer and thus improving the frequency response of the bridge.

The terminating resistors R 111 and R 112 are also built up in a balanced condition to suppress any field unbalance and to ensure a low-inductive termination.

The bridge load impedance  $Z_l$  is essentially formed by the relatively high impedance resistors R 109 and R 110 so that the bridge load impedance is high compared to its parallel impedance ( $\approx R$ , see Fig. 20).

### 2.6.3. Reflection Standards

Standard resistors N with socket and plug connection together with calibrating resistors  $R_{cal}$  are provided for the various impedance values of the PSM-5. The calibrating resistors cause a reflection coefficient  $r = \frac{R_{cal} - N}{R_{cal} + N} = \pm 0.1$ , corresponding to a reflection coefficient of 20 dB (2.3 Np).

The sign has been chosen so that the calibrating resistors are within the ranges 50  $\Omega$  to 75  $\Omega$  or 100  $\Omega$  to 150  $\Omega$ .

Z / $\Omega$	50	60	65	75	135	150
r	+ 0.1	+ 0.1	- 0.1	- 0.1	- 0.1	- 0.1
$R_{cal} / \Omega$	61.11	73.33	53.18	61.36	102.27	122.73

The parallel capacity, which results from the correct coaxial installation of the resistors, is compensated for by a series inductance.

### 2.6.4. Measuring Cable

A resistive transformation and thus a transformation of the reflection coefficient will take place, if the impedance of the measuring cable deviates from the necessary Z value. The maximum error thus occurring would be at a cable length of  $\lambda/8$ .

The 25 cm measuring cables are, however, sufficiently short and allow, in spite of this, a rational connection of the test object. Since it is necessary to use a 60  $\Omega$  cable for  $Z = 65$ , the error so caused will increase the reflection attenuation. Cable or adapter errors will add themselves to the error of the test object.

3.

### LIST OF THE REQUIRED MEASURING INSTRUMENTS for fault finding and testing of the units

## a) Common Instruments

Instrument	Required Measuring or Frequency Range	Required Accuracy	Recommended Instruments	Manufacturer
DC Multimeter	0 to 200 V 0 to 6 A	1 %	UM, 100 k $\Omega$ /V	AEG Gossen
Ohmmeter	0 to 150 $\Omega$		Triohm	Gossen
Oscilloscope with 10:1 probe	0 to 33 MHz	Risetime < 10 ns	OS-1 TK-4	W.u.G. W.u.G.
RF Millivoltmeter	500 kHz to 105 MHz	$\pm 3\%$ f.s.d.	411 A	hp
Frequency Counter	10 Hz to 36 MHz		FZ-2	W.u.G.
Signal Generator	10 Hz to 30 MHz		MG-64	W.u.G.
VHF Power Generator with Plug-In Unit	0 dB 30 to 90 MHz 85 to 175 MHz		LMS-68 LO-30 LO-85	W.u.G.
Calibrated Attenuator	0 to 120 dB resp. 0 to 12 Np		FE-1	W.u.G.
Wideband Voltmeter with probe	10 Hz to 600 kHz	$\pm 3\%$ f.s.d.	SM-1 TK-1	W.u.G.
Selective Level Meter	10 kHz to 36 MHz		PSM-5	W.u.G.
Display Unit			SG-1	W.u.G.
Plug-In Unit			SGE-10	W.u.G.
Probe	10 kHz to 36 MHz		TK-8	W.u.G.
DC Power Supply	0 to 15 V		Konstanter	Gossen
DC Power Supply	0 to 36 V		6960 A	hp

## 4. FAULT FINDING

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### 4.1. Oscillator Portion RWO-5

#### 4.1.1. Cabinet Removal and Access to the Sub-Units

The oscillator unit must be removed from its cabinet during the fault finding procedure. This is achieved by releasing the side-mounted holding screws and by removing the unit from the cabinet by means of the two handles.

**Caution!** Disconnect mains plug previous to removing the cabinet!

Stand the unit on the power supply side, the hinged chassis can now be opened by releasing two screws (see Fig. 24)

The following sub-units are now accessible:

Coarse frequency oscillator with auxiliary oscillator

Interpolation oscillator

Dial discs of the interpolation and coarse frequency oscillator (after removing the covering plate)

Amplifier and counter discriminator I

Counter discriminator II

The two 10 : 1 frequency dividers

The last four sub-units are plug-in boards and can be removed after releasing the holding screws.

Lock-in indication amplifier

This sub-unit is accessible after removing the 4 plug-in boards.

**Caution!** Both counter discriminators are fed from a common zener voltage. The zener diode is to be found in counter discriminator II. This means that counter discriminator I cannot be used without counter discriminator II.

Both 10 : 1 frequency dividers (12a) and (12b) are supplied with voltage via a common voltage dropper. Thus operation with only one board is not possible.

Since the two cards are completely identical, they may be exchanged.



#### 4.1.2. Faults in the Power Supply

A fault can be assumed to be present in the power supply if after switching on, the projection dials remain unilluminated and no meter indication is observed with level control P 1002 fully clockwise.

Firstly it should be established if the correct mains voltage is provided. After this the fuses Si 2402, during mains operation, and Si 2401 during battery operation should be checked for continuity. In addition to this it is also necessary to examine the polarity of the battery during battery operation (Plus to point 4 or point 5 of socket Bu 2401). The following voltages should be measured against ground at the various test points on circuit board 241-AX1 "additional filtering" (rear of RWO-5), with a 100 k $\Omega$ /V DC voltmeter (Point 1 = ground):

Point	Nominal voltage
6	+180 V
5	+ 36 V
3	+6.3 V
2	- 3 V
4	- 18 V

The plug-in power supply connections from receiver or generator should be removed if large deviations from the above nominal values are determined. If the given values are now indicated, this would indicate a fault in either generator or receiver. If however the deviation from the nominal value remains, then the fault will be within the power supply or in the Oscillator RWO-5.

A short circuit is indicated by an oscillation in the power supply.

(For details regarding the repair, please see the section "Checking and Alignment").

#### 4.1.3. Malfunction of the Coarse Frequency Oscillator Locking indicator

- Symptoms:
- No indication at Meter J 2302 on tuning the coarse frequency oscillator (Switch S 401 in "LOCKED" position).
  - Needle returns to the 1 MHz portion of the meter when a 100 kHz frequency locking has been selected.

#### 4.1.4. Malfunction of the Interpolation Oscillator Tuning Indicator

Symptom: No deflection at meter J 2301 on tuning the interpolation oscillator (S 1602 to position 1).

Possible causes: Non-operation of the 10 kHz reference frequency.

Non-operation of the Interpolation oscillator frequency.

Checking of meter J 2301

Switch sweep width S 1602 to position 3: the pointer of meter J 2301 should now indicate f.s.d. Return switch to position 1.

Connect an oscilloscope via a 10 : 1 probe to point 1 of the tuning indicator (19) 241-AA (top of RWO-5).

Oscillogram 10 should be displayed.

It is necessary to remove the generator and receiver from the cabinet during the following measurement. The generator and receiver portions can be removed as individual units after releasing the carrier, synchronizing and power supply cables and the mechanical interconnections.

After this, connect an oscilloscope to point 2 of the tuning oscillator and set the interpolation oscillator to the 0 dial marker.

Oscillogram 11 should now be displayed.

If both signals are correct, then the fault is to be found in the tuning indicator itself or in the subsequent discriminator (23) .

The tuning indicator can be assumed to be operating correctly if oscillogram 12 can be displayed at the collector of T 1906 and if a periodic frequency minimum can be noticed at point 5.

Printed circuit board "Discriminator" (23) 241-G is to be found in the upper portion of the cast cabinet of the interpolation oscillator below the "tuning indicator" circuit (top of RWO-5). This circuit is accessible after removing the screening of the "tuning indicator" by releasing the connection cables and the side-mounted holding screws, and removing the screening in an upwards direction. The discriminator indicator circuit is mounted directly at the rear of the locking indicator meter J 2301. The plug connections are accessible from the base of the RWO-5.

- b) Measure the voltage at point 2 of the frequency and phase control with a 100 k $\Omega$ /V DC multimeter. The reading must amount to +36 V if relay 801 has operated. Check contact rel 803 with a continuity meter.
- c) The principle of operation of the unit requires that the difference between the frequencies of the control and coarse tuning oscillators always coincide to the frequency of the interpolation oscillator. It is thus necessary that the input frequencies of the two counter discriminators of an instrument also coincide.

Checking the generation of the interpolation frequency:

Measure the input signal of counter discriminator II at point 4 of the buffer and switching stage (21) with an oscilloscope via a 10 : 1 probe.

Place switch S 1602 to position 1 and set interpolation oscillator to dial marker 100. Oscillogram 15 should now be displayed (see Section 6.1.)

Interpolation frequency range "x 10" :

Place switch S 1602 to position 2.

The switching stage is fed in this position with the unconverted frequency of the interpolation oscillator. Oscillogram 16 should now be displayed. If the signal exhibits the correct curve and amplitude, then the interpolation oscillator is operating correctly. The fault must therefore be in one of the sub-units: buffer and switching stage; frequency converter; 10 : 1 divider or 2.7 MHz oscillator.

Fault finding by signal tracing:

Take an oscillogram at point 2 of the 400 kHz low-pass filter (20) (241-D).

Oscillogram 13 should be displayed (see Section 6.1).

The 2.7 MHz signal, which is fed to the frequency converter (241-E), should now be checked. However, it is necessary to firstly disconnect the input of the converter (point 3) to avoid any retro-effect from the mixer. Oscillogram 14 should be displayed. Reconnect the input connection.

The oscillogram at the collector of T 2102 should represent a sine wave signal of  $U_{ss} = 1.8 \text{ V}$ .

## 4.2. Generator PS-5

### 4.2.1. Complete Failure of the Unit

Symptoms: No meter deflection

No output level

Possible causes: Absence of the operating voltages.

Non-operation of the carrier frequency supply

Fault within the combination 69 MHz oscillator,  
amplitude modulator, 69 MHz amplifier

Fault within the combination of pre-amplifier and  
final amplifier.

### 4.2 .1. Checking the Operating Voltages

Measure the operating voltages at points 6 and 8 of the voltage filter network (12) (242-S) with a DC multimeter.

Nominal values: -18 V

+36 V

If the nominal values are obtained, it is then necessary to measure the total current consumption by placing ammeters into the leads to points 6 and 8 of the voltage filter network.

Nominal values: Point 6 : 390 mA

Point 8 : 120 mA

If these nominal values are not obtained, the fault can be localized by measuring the current consumption of the individual sub-units and by comparing them to the table given in Section "Checking and Alignment".

### 4.2.1.2. Checking of the Carrier Frequency Supply

The control frequency originating from the RWO-5, known here as the carrier frequency, is checked by measuring the input voltage at test point 3 (TP 3) of the limiter (11) (242-P). This should amount to 500 mV with the coarse tuning oscillator of the RWO-5 set at 0 MHz.

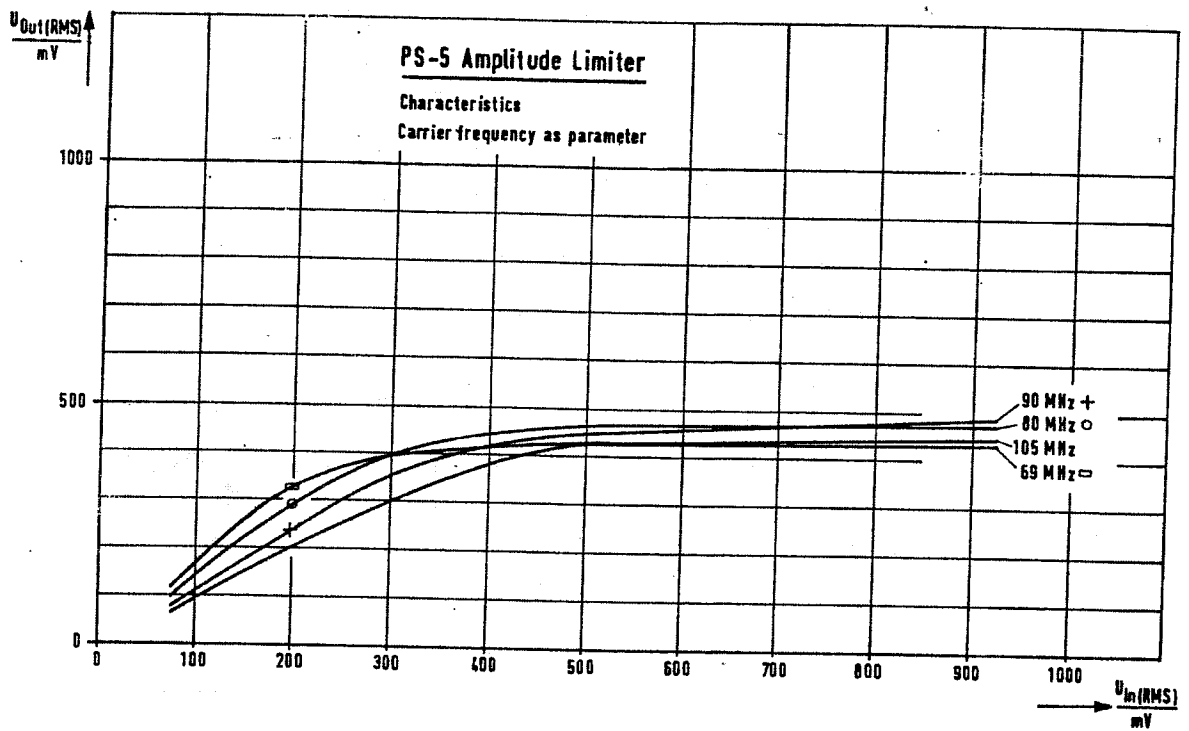


Fig. 26

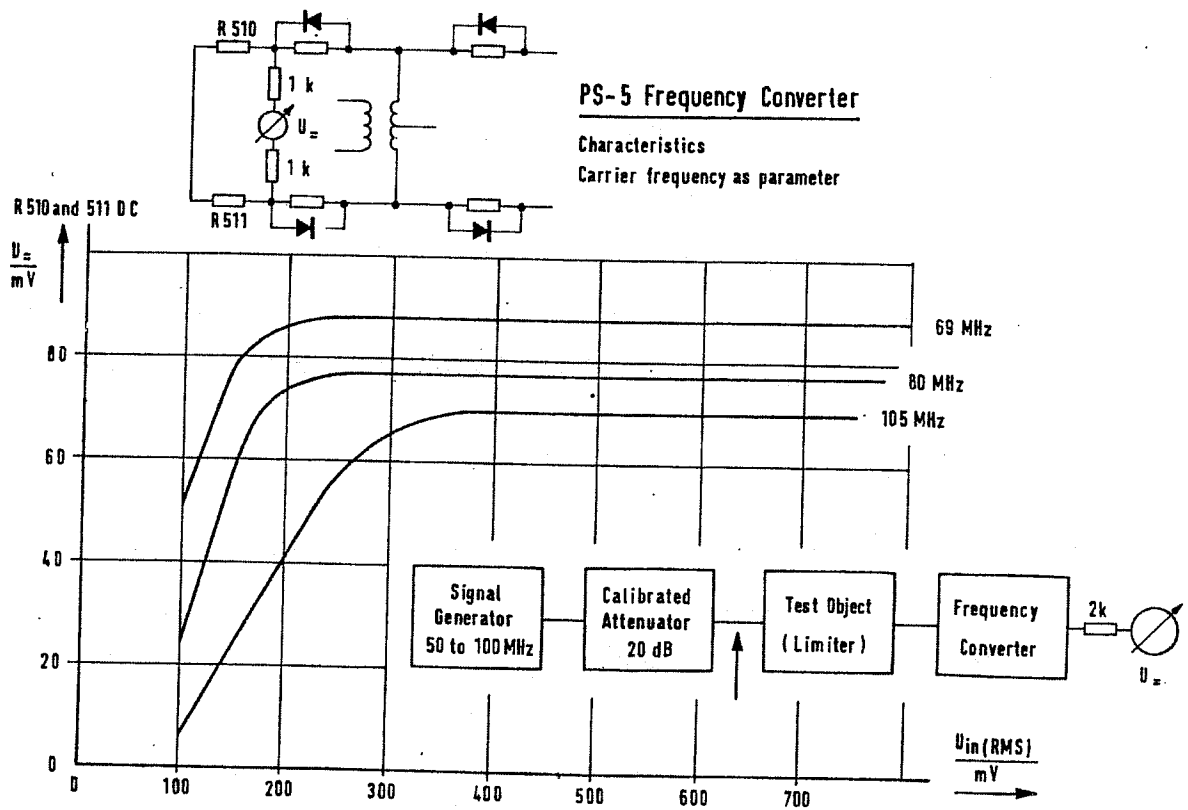


Fig. 27

If the sideband frequency at the output of the low-pass filter also returns to 0 in the zero setting of the two RWO-5 oscillators, it can be assumed that the combination of 69 MHz oscillator, amplitude modulator and 69 MHz amplifier is operating correctly. If no signal is obtained at the output of the unit (output II) with the correct sideband signal at the low-pass output, the fault should be found in the combination of pre-amplifier or final amplifier.

If the sideband signal is not present, the fault finding should be concentrated to the combination 69 MHz oscillator, amplitude modulator and 69 MHz amplifier.

#### 4.2.4. Malfunction in the Combination 69 MHz Oscillator, Amplitude Modulator and 69 MHz Amplifier

##### 4.2.4.1. Checking the 69 MHz Oscillator

Measure the output level of the 69 MHz oscillator (1) with a RF millivoltmeter at test point 1 (circuit board 242-R)

Nominal value:  $\geq 1.7$  V

##### 4.2.4.2. Checking the 69 MHz amplifier

Measure the input level of the 69 MHz amplifier (4) at test point 1 (circuit board 242-B), whereby the level potentiometer should be fully advanced.

Nominal value:  $> 50$  mV.

Reduce the input level with P 1002 to 20 mV and measure the output level of the 69 MHz amplifier at test point 3.

Nominal value: 110 mV.

##### 4.2.4.3. Instructions for Fault Finding in the 69 MHz Amplifier

A DC measurement cannot be used to establish whether one of the two transistors, T 402 or T 403, are defective.

It is possible, due to the different characteristic curves, that in the absence of an AC input signal one transistor takes over the whole current, whereas the other is blocked. Therefore, if one of the two emitter voltages is equal to the voltage at the connection of R 409 and R 410, a further test should be made to establish whether an unfavourable pairing prevails.

#### 4.2.5. Malfunction in the Combination of Pre-amplifier and Final amplifier

This combination can be checked by carrying out an amplification measurement.

Adjustment at the RWO-5: Interpolation oscillator to 0 kHz

Coarse frequency oscillator to 1 MHz

##### 4.2.5.1. Amplification of the Pre-amplifier

Adjust the voltage at test point 1 of the pre-amplifier (7) (circuit board 242-F) with potentiometer P 1002 to 10 mV.

Measure the voltage at test point 2 of the pre-amplifier.

Nominal value: 282 mV

##### 4.2.5.2. Amplification of the Final Amplifier

Adjust the voltage at test point 2 of the pre-amplifier with Potentiometer P 1002 to 100 mV.

Terminate socket Bu 802 with 75  $\Omega$ .

Measure the voltage at Bu 802 (Output II).

Nominal value: 325 mV

Additional values of interest during fault finding:

Pre-amplifier

Amplification at low frequencies: 26.9 to 30.9 dB; average: 28.9 dB

Final amplifier

Amplification at low frequencies (Bu 802 terminated with 75  $\Omega$ ):

10 to 10.6 dB; average: 10.3 dB

Recommended input voltage to avoid over driving: 100 mV

The following curve represents the frequency response of the final amplifier:

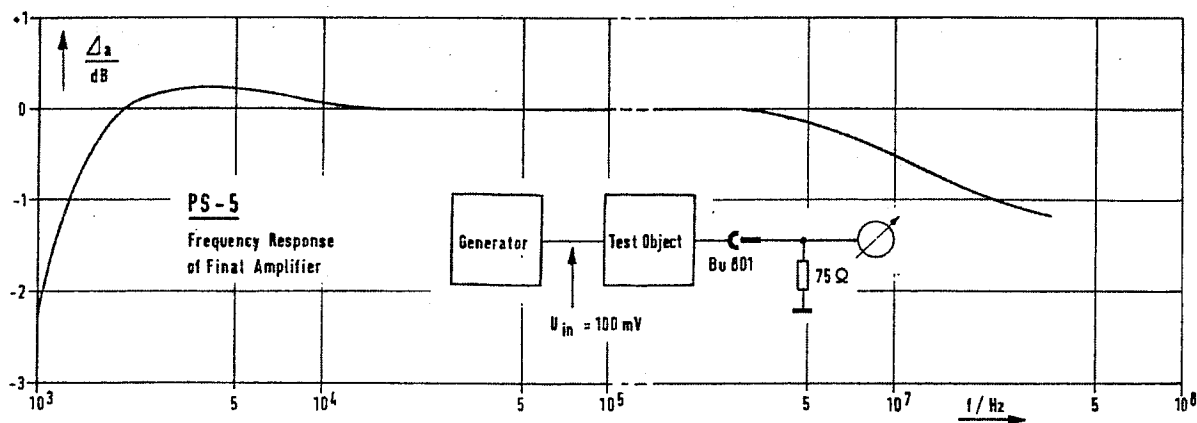


Fig. 31

Connect an oscilloscope via a 10 : 1 probe to the following points and compare with the relative oscillograms in Section 6:

Collector of T 202: Oscillogram 20

Primary of  $\dot{U}$  201, between 1 and ground: Oscillogram 21

Secondary of  $\dot{U}$  201, between 6 and ground: Oscillogram 22

Secondary of  $\dot{U}$  201, between 3 and ground: Oscillogram 23

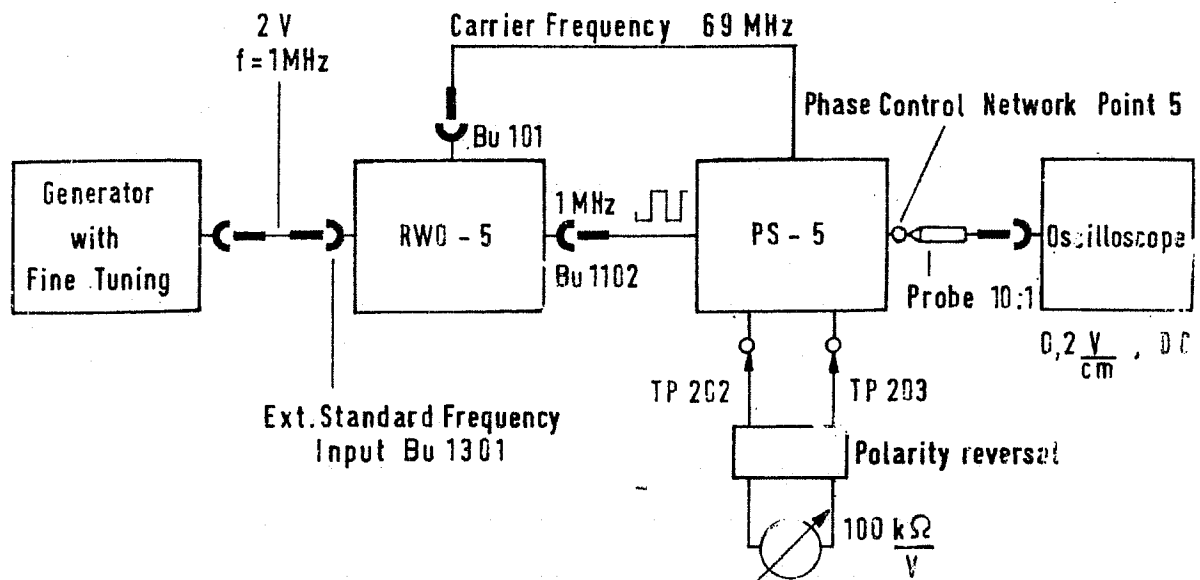


Fig. 32 Measuring arrangement

The external generator frequency of 1 MHz replaces the internal 1 MHz standard frequency. This is possible because the internal 1 MHz standard frequency of the RWO-5 is switched off on connecting a generator to socket Bu 1301.

It is then possible to alter the 1 MHz synchronizing frequency for the generator with the fine tuning and to thus measure the hold and pull-in range of the generator phase control.

The hunting oscillator is non-operative within the hold range and the voltmeter and oscilloscope will indicate a DC voltage which increases with the advancing detuning from 1 MHz.

Since the DC voltage can be of negative or positive polarity, a polarity reversal has been placed before the meter.



Switch S 1001 to "VOLTAGE LEVEL"

Disconnect the carrier signal by removing St 1101 from socket 101 of the RWO-5.

The output level should now return to zero.

The meter needle should be directly at the mechanical zero point (correct if necessary with P 1004).<sup>1)</sup>

Reconnect carrier signal.

Measure the voltage at socket Bu 802 with a RF millivoltmeter and adjust to 1.55 V with P 1002.

The built-in meter should now show a reading of 0 dB (0 Np) (correct if necessary with P 802).

#### 1.2.8. Malfunction of the Amplitude Control

Symptoms: The meter indicates a level reduction of 4.5 dB (0.5 Np), on placing the " $R_i / \Omega$ " switch S 902 from 50  $\Omega$  to 150  $\Omega$  in the "POWER LEVEL" mode, whereas the output level remains constant.

Possible causes: The control stage is open circuit (socket Bu 1001 is defective).

##### 1.2.8.1. Operational Check of the Amplitude Control

The meter reading should be adjustable between approximately zero and f.s.d. with level potentiometer P 1002.

Place switch S 1001 to "POWER LEVEL" and " $R_i / \Omega$ "

Switch S 902 to 150  $\Omega$ .

Measure the output amplitude at socket Bu 802 at a meter reading of 0 dB (0 Np) with a RF millivoltmeter.

Nominal value 2.45 V (2.11 V)

The output amplitude must be progressively lower at each position on switching from  $R_i = 150 \Omega$  to  $R_i = 50 \Omega$  and should have a value of 1.42 V (1.22 V) at  $R_i = 50 \Omega$ , if 0 dB (0 Np) has been reset.

The reading on the built-in meter should, however, only increase by approximately 0.7 dB (0.07 Np).

---

1) Commencing with series G (or series F if J 1001 accords to Bv 242-8104 and 242-8105) :

Connect 100 k $\Omega$ /V DC voltmeter parallel to R 1016. Select 300 mV range and adjust P 1004 until 75 mV reading is obtained. Caution! The reading should increase on clockwise rotation of P 1004.

Attenuator position	20 dB pad (2 Np pad)	10 dB pad (2 Np pad)	20 dB pad (2 Np pad)	20 dB pad (2 Np pad)
0 dB (0 Np)	rel 901	rel 903	rel 905	rel 907
-10 dB (-1 Np)	rel 901	rel 904	rel 905	rel 907
-20 dB (-2 Np)	rel 902	rel 903	rel 905	rel 907
-30 dB (-3 Np)	rel 902	rel 904	rel 905	rel 907
-40 dB (-4 Np)	rel 902	rel 903	rel 906	rel 907
-50 dB (-5 Np)	rel 902	rel 904	rel 906	rel 907
-60 dB (-6 Np)	rel 902	rel 903	rel 906	rel 908
-70 dB (-7 Np)	rel 902	rel 904	rel 906	rel 908

Measure the voltage across the coils of the relay that should be energized in the malfunctioning position.

If the voltages are present, the contacts should be checked for continuity.

If a voltage is absent, this will indicate a defective diode.

The following table shows the diodes which are in circuit in the individual attenuator positions.

Attenuator position	GI 901	GI 902	GI 903	GI 904	GI 905	GI 906	GI 907	GI 908	GI 909	GI 910
0 dB				X				X		
-10 dB				X				X		
-20 dB	X				X				X	
-30 dB	X				X				X	
-40 dB		X				X				X
-50 dB		X				X				X
-60 dB			X				X			
-70 dB			X				X			

Schaltbuchse Bu 1001  
Contact socket

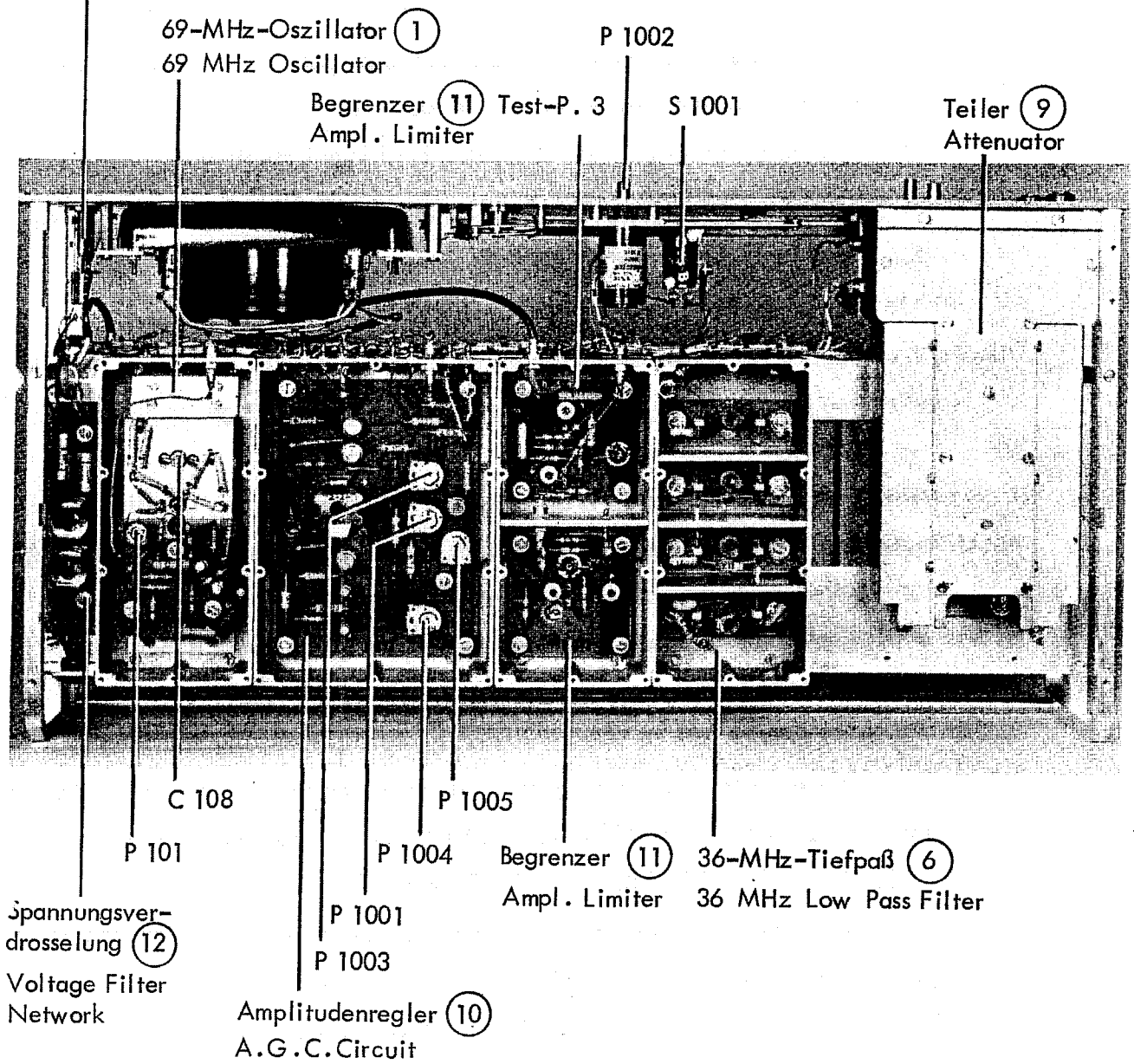


Bild 36 PS-5 Ansicht von unten  
Fig. 36 PS-5 Bottom View

Eingangsbaustein  
Input Network

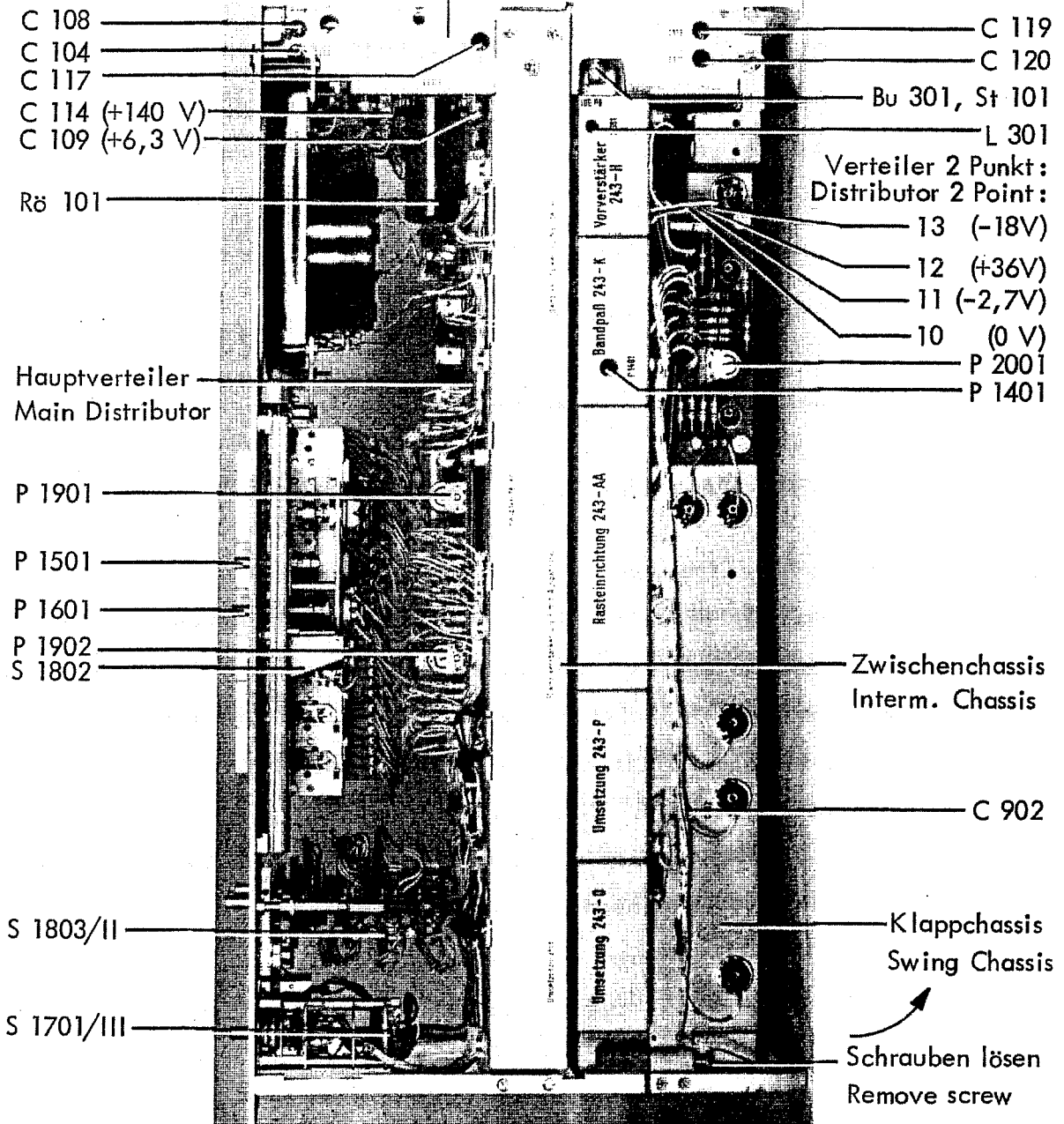


Bild 37 PM-5 Ansicht von oben  
Fig. 37 PM-5 Top View

Because of this the test points were placed within the bandpass filters. Small test points have been provided for frequencies in the highest IF range of 69 MHz, because the possibility of spurious signal introduction is extremely great. A number of test points have been screened by a small metal cap for the same reason.

A list of the test points used in the wideband measurement is given on page 117. The measurement is made in the order given. The hinged chassis is now tilted out after loosening the two screws and the connection cable at the rear of the PM-5 placed between plug St 101 and socket Bu 301.

The given levels are valid for the linear calibration position and a deviation of 20 % is allowed.

Test points 2, 3, 8 and 9 are potentials isolated from ground and points 13, 14 and 15 indicate the locked condition of the carrier oscillators.

After the defective sub-unit has been found, a thorough examination can be made by checking the operating points against the values given in the circuit diagram, by measuring the gain and loss values according to the level chart, additionally testing the operation of the relays with the relay plan and viewing the signal waveforms by means of an oscilloscope etc.

The level chart on page 119 gives a list of the level values of the individual sub-units in the linear calibration position. These levels are, with the exception of the carrier level, selectively measured and the corresponding level meter connected via a high-impedance probe.

#### 4.3.2.2. Checking the Frequency Locking

The internal carrier oscillators of the PM-5 are locked to each other and to a 10 kHz standard frequency spectrum of the RWO-5. This guarantees the high frequency accuracy and stability of the tuning and frequency tracking of generator and receiver.

The 61.56 MHz oscillator operates more than 270 kHz from its nominal frequency in its unlocked condition, so that no calibration signal will be present on tuning to 100 kHz. Furthermore, the generator output voltage will not be indicated in the receiver portion during operation as PSM-5.

The oscilloscope will display the following conditions:

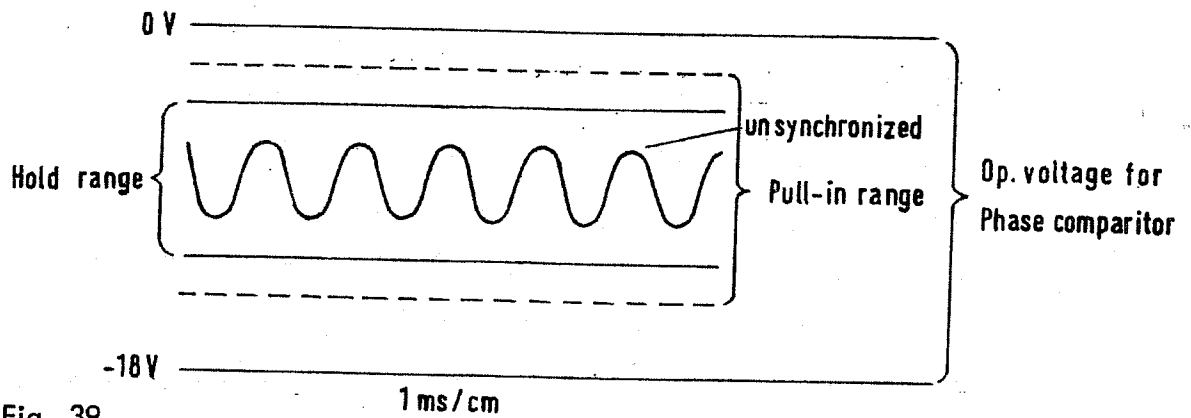


Fig. 39

The harmonic of the synchronizing frequency should be in centre of the hold range. The hold range can be measured by injecting an external synchronizing frequency, the exact frequency of which is indicated on a connected frequency counter. The oscillator frequency and hold range are then calculated by multiplication. The oscillator should be tuned so that the resulting frequency is symmetrical to the hold range. The control circuit should not oscillate within the hold range. The control DC voltage should not possess any AC component.

Measurement values for the three carrier oscillators:

No	Oscillator	$f_{\text{sync}}/\text{MHz}$	Synchronizing voltage $U_{\text{pp}}/\text{V}$	Injection point	Hold range	Oscillator alignment
13	570 kHz	1.0	2	RWO-5 1) Bu 1301	$\pm 28.5$ Hz	Ü 1203
14	6.84 MHz	1.71	3	243-0, Pkt.5 2) via 0.1 $\mu\text{F}$	$\pm 36$ kHz	Ü 1103
15	61.56 MHz	20.52	2.5	243-U, Pkt. 1 2)	$\pm 270$ kHz	C 902

- 1) The output impedance of the generator must have a maximum galvanic connection of 100 k $\Omega$  to ground.
- 2) Remove connection to adjacent sub-unit.

The hold and pull-in range of the 570 kHz crystal oscillators are identical.

The Level Measuring Set PSM-5 allows an easy means of checking the frequency locking by feeding a signal from the PS-5 to the PM-5. If the PM-5 is in the bandwidth position "500 Hz", the same level should be indicated at generator and receiver.

#### 4.3.3.1. Sensitivity Selection

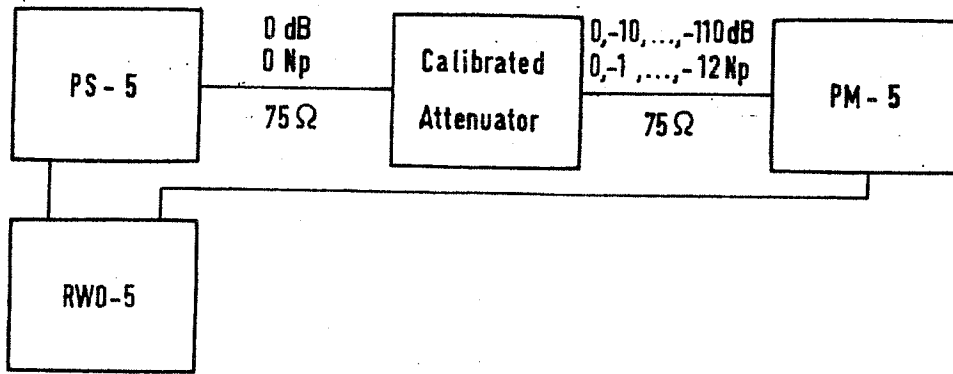


Fig. 40

The level values are adjusted with the attenuator and all sensitivity positions "LOW NOISE" and "LOW DISTORTION" must be checked both in the linear and logarithmic mode.

#### 4.3.3.2. Linear and Logarithmic Calibration

Tune RWO-5 to 100 kHz. Relay Rel 112 will be energized by depressing the calibration button.

#### 4.3.3.3. Bandwidth

Select linear calibration position; lock coarse frequency oscillator to 100 kHz; sweep width selector to position "0...100 kHz"; measure the 3 dB (0.35 Np) bandwidth by tuning the interpolation oscillator and observing the PM-5 meter.

Nominal values: broad  $f = \pm 1.75$  kHz

narrow  $= \pm 250$  Hz

The relay voltage is fed via distributor II (243-BB) point 11.

#### 4.3.3.4. Demodulation Modes

Check the converter frequencies with a Frequency Counter FZ-1 connected via a 0.5 μF capacitor to either S 1701/III or 243-A point 1.

Demodulator	Bandwidth	Frequency
S 1701	S 1801/6	f/kHz
	broad	28
	narrow	29
	narrow	31
	broad	32

### 4.3.5. Frequency Response

#### 4.3.5.1. $Z = 75 \Omega$

The basic frequency response of the PM-5 is compensated for in the pre-amplifier by a frequency dependent negative feedback.

The frequency response of various pre-attenuator positions can be coincided with another. For the frequency response alignment, the difference between the input and output voltage is observed. The input voltage is fed via a  $75 \Omega$  resistor to the receiver input, in order to ensure that the  $75 \Omega$  input impedance of the PM-5 is also taken into consideration. The generator voltage is measured with an RF voltmeter possessing a DC output. The level of the PS-5 should not be in advance of  $\approx -10$  dB, so that a low distortion factor is maintained ( $< 1\%$ ). If the Sweep Measuring Set WM-50 is available, the frequency response curves can be displayed.

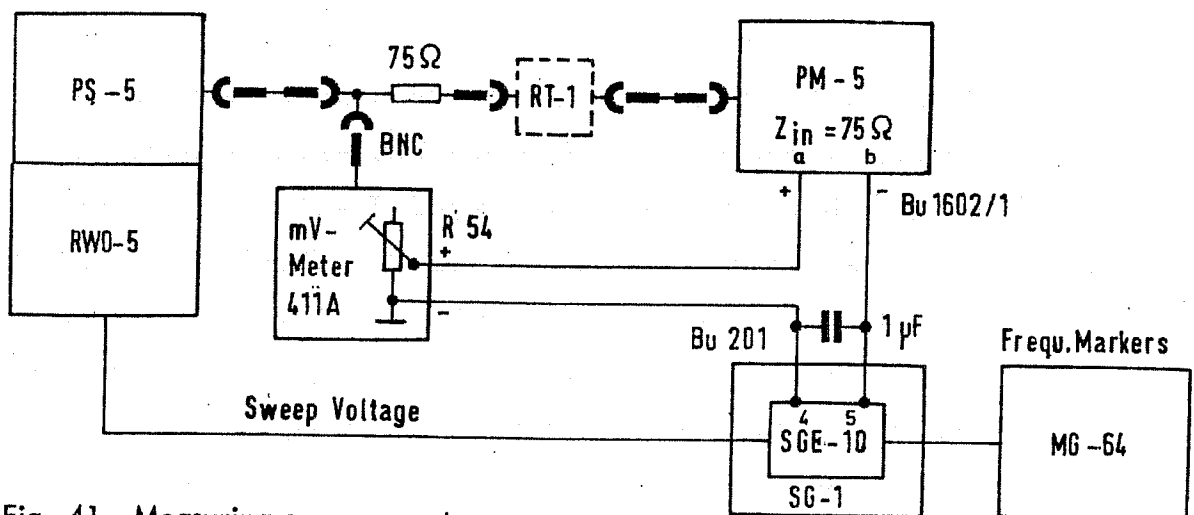


Fig. 41 Measuring arrangement

Generator portion PS-5: Select 0 dB (0 Np)  $Z_{out} = 75 \Omega$ , output level  $\approx -10$  dB ( $-1.15$  Np), ie.  $\approx 310$  mV reading. Connect PS-5 via a  $75 \Omega$  standard resistor, without RT-1, to the PM-5. The reflection coefficient ( $r$ ) of the resistor should be less than 0.005 at 36 MHz.

The measuring point for the millivolt meter must be directly adjacent to the resistor. The most favourable form would be a T-piece with built-in resistor and suitable connections.

Receive portion PM-5:  $Z_{in} = 75 \Omega$ , level control to  $-10$  dB ( $-2$  Np), linear and low noise modes, set calibration adjustment for meter f.s.d.



The frequency response can only be made in a point-for-point manner if the level meter does not possess a DC output. The input level must be adjusted for a constant reading at each measuring position.

The frequency response can also be checked, with a lower degree of accuracy, by directly connecting the generator output and receiver input.

The stabilized generator output supplies good results, especially at low frequencies, where the built-in characteristic impedances are still real, and this method can be used for check purposes over the whole frequency range.

The error consists of the frequency response error of the generator level (frequency response of the DC control voltage) and the dividing error between  $Z_{out}$  of the PS-5 and  $Z_{in}$  of the PM-5.

#### 4.3 .2. $Z = 50, 60, 65, 135$ and $150 \Omega$

The frequency response of the receiver portion has been optimally aligned for an impedance of  $75 \Omega$ . The dividing error between  $R_N$  and  $Z_{in}$  has been taken into consideration by feeding via a  $75 \Omega$  standard resistor.

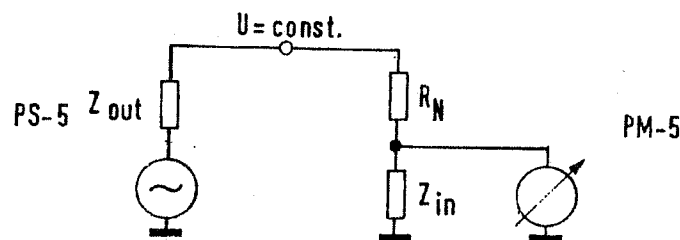


Fig. 42

Frequency response for impedances of  $50 \Omega$ ,  $60 \Omega$  and  $65 \Omega$ :

These impedance values are obtained by placing additional resistor values in parallel. The reflection coefficient varies only slightly from the  $75 \Omega$  position and the characteristic impedance curve is similar.

The frequency response is therefore hardly effected and is a maximum of 1 % better or worse up to 15 MHz and 2 % better or worse up to 36 MHz.

Frequency response for impedances 135  $\Omega$  and 150  $\Omega$ :

The input capacity is neutralized in position  $Z_{in} = 150 \Omega$ .  $Z_{in} = 135 \Omega$  is also formed by parallel resistor switching. The reflection factor and the transfer locus of the characteristic impedance are virtually identical for both impedance values. The frequency response error possesses, in the most unfavourable condition, the sum of the two reflection coefficients for  $75 \Omega$  and  $150 \Omega$ . The maximum frequency response error for  $Z_{in} = 75 \Omega$  is less than 2 % up to 15 MHz.

#### 4.3.6.2. IF Suppression

The attenuation of the 1. IF of 69 MHz is essentially determined by the 69 MHz attenuation peak of the 39 MHz low-pass filter. See 5.3.6. for details regarding measurement and alignment of C 406.

Nominal value for all tuning frequencies  $a \cong 70$  dB (8 Np).

#### 4.3.6.3. Image Rejection

The image rejection for mixer I can be measured as described in 5.3.6. for input frequencies between 138 and 174 MHz, corresponding to a tuning frequency of 0 to 36 MHz.

Nominal value:  $a \cong 70$  dB (8 Np)

The image rejection values for the other mixers are greater than this. The input signal of 0 dB is supplied from a Signal Generator MG-64 and the receive level then directly corresponds to the image rejection.

Nominal value:  $a \cong 80$  dB (9 Np)

Input frequency	Tuning frequency	Mixer
160 kHz	100 kHz	IV
2.2 MHz	1 MHz	III
18.88 MHz	4 MHz	II

#### 4.3.7. Disturbing Signals

##### 4.3.7.1. Intrinsic Noise

The intrinsic noise of the receiver is constant at tuning frequencies above 1 MHz. The noise level increases at lower frequencies. The values are dependent on the carrier balance of the mixers and on the noise characteristics of the mixer diodes GI 501 to GI 504. The noise level in the 500 Hz bandwidth position is 9 dB (1 Np) lower than in the 3.5 kHz position.

The measurement is carried out with a calibrated PM-5 in its most sensitive position of -110 dB (-12 Np); the input socket open; input impedance  $Z_{in} = 75 \Omega$  and 3.5 kHz bandwidth.

Nominal value:    Tuning frequency    Meter reading  
   80 kHz                     $\cong -11$  dB (-1.5 Np)

List of possible interfering signals:

Tuning frequency in kHz	Cause of interference
10	10 kHz standard frequency
10.5	Tube microphonics
approx. 18	Two point regulator frequency
approx. 28	Chopper frequency
80	Noise $\leq -121$ dB ( $\leq -13.5$ Np)
approx. 56	2 x chopper frequency
63.9 $\pm$ 2.5 kHz	Auxiliary oscillator frequency
approx. 84	3 x chopper frequency
100	100 kHz standard frequency
approx. 112	4 x chopper frequency
approx. 128	2 x auxiliary oscillator frequency
approx. 192	3 x auxiliary oscillator frequency
222.6	9 x carrier I - 10 x carrier II
570	570 kHz oscillator frequency
1000	1 MHz standard frequency
1710	3 x 570 kHz oscillator frequency
3000 } 6000 }	Coarse or interpolation oscillator frequency
6090	5 x carrier II - 4 x carrier I
8880	8 x carrier I - 10 x carrier II
11210	7 x carrier I - 9 x carrier II
15560	3 x carrier I - 4 x carrier II
19620	3 x carrier II - 2 x carrier I
20560	3 x 6.84 MHz oscillator frequency
27060	2 x carrier I - 3 x carrier II
36080	3 x carrier I - 5 x carrier II

The following oscillogram was obtained on a display unit with the Sweep Measuring Set WM-50 at a sweep width of  $\pm 35$  kHz; the MG-64 at 2.5 MHz; the PM-5 in log. position and at a bandwidth of 3.5 kHz.

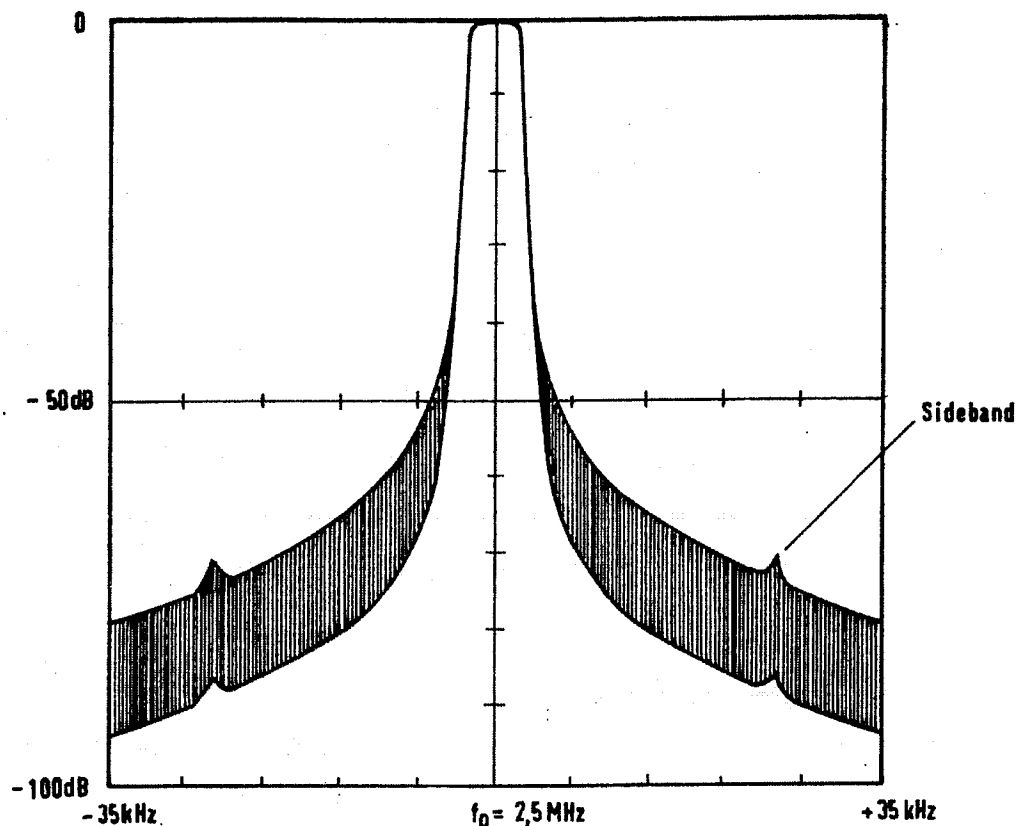


Fig. 43

AC line interference can be avoided by battery operation. The power transformer in the RWO-5 is mounted on felt supports to keep vibrations at a minimum. Such vibrations could cause an unwanted deviation of the interpolation oscillator and care should also be taken when other equipment is to be operated on the same table etc. In the Measuring Set PSM-5, however, the generator and tuning frequency will vary synchronously.

#### 4.3.8. Transfer Impedance

The ground connection between generator and test object must be low-impedance when making measurements on high loss test objects. This is to ensure that the whole generator current flows back via this link. The ground of the PM-5 is isolated, in respect to the measuring frequencies, from the grounds of the RWO-5 and PS-5, and also from the cabinet.

Test points

PM-5 to "CAL" linear; calibration control to 0 dB (0 Np); RWO-5 to 100 kHz.

Necessary measuring equipment: DC voltmeter 100 k $\Omega$ /V (Gossen)

Wideband voltmeter SM-1 with probe TK-1 (W u.G)

RF millivoltmeter type 411 A (hp)

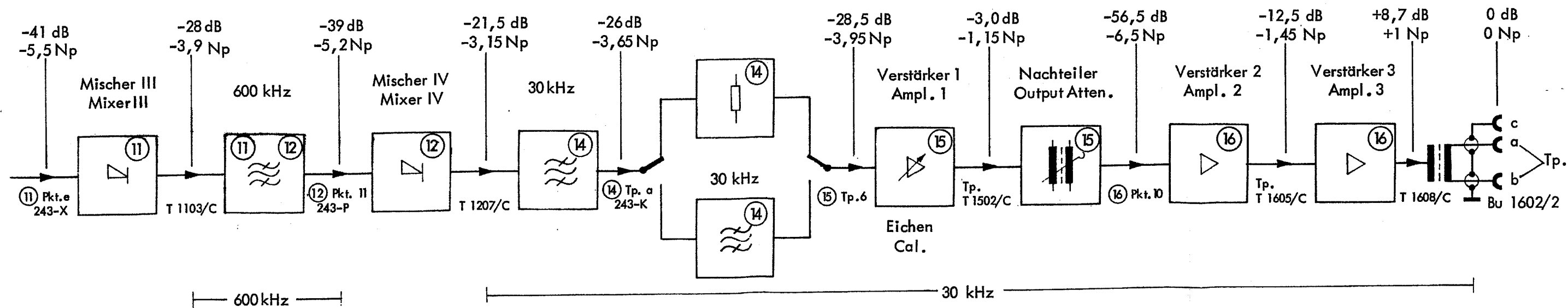
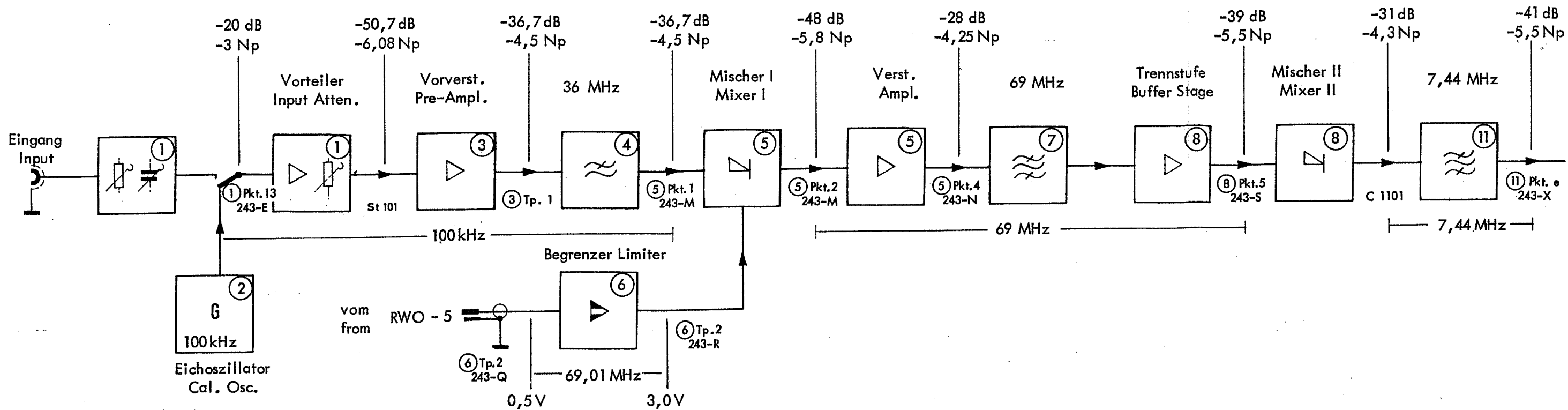
All voltages are measured against ground

No.	Test point	Frequency f/MHz	Measuring equipment	Voltage at the test points		Notes
				dB version	Np version	
1	243-E Tp.4	0.10	SM-1	-20 dB	-3 Np	
2	243-E Tp.7	0.10	SM-1	-50.7 dB	-6.08 Np	1) 2)
3	243-H Tp.1	0.10	SM-1	-36 dB	-4.5 Np	1) 2)
4	243-Q Tp.2	69.10	411 A	$\approx$ 0.50 V	0.50 V	2)
5	243-R Tp.2	69.10	411 A	$\approx$ 3.0 V	3.0 V	2)
6	243-W Tp.a	7.44	411 A	52 mV	58 mV	2)
7	243-Z Tp.a	0.60	SM-1	-26 dB	-3.75 Np	2)
8	243-K Tp.a	0.03	SM-1	-25 dB	-3.6 Np	1)
9	243-C Tp.6	0.03	SM-1	-27 dB	-3.85 Np	1)
10	243-C Tp. C	0.03	SM-1	-1.5 dB	-0.95 Np	Casing T 1502
11	243-D Tp. C	0.03	SM-1	-12.6 dB	-1.45 Np	Casing T 1605
12	Bu 1602/2 Tp.	0.03	SM-1	0 dB	0 Np	between a and b
13	243-AA Tp.5	-	100 k $\Omega$ /V	$\approx$ -9.5 V; R <sub>p</sub> .= 330 k $\Omega$		2) 3)
14	243-Y Tp.4	-	100 k $\Omega$ /V	$\approx$ -5.5 V; R <sub>p</sub> .= 68 k $\Omega$		3)
15	243-V Tp.3	-	100 k $\Omega$ /V	$\approx$ -5.5 V; R <sub>p</sub> .= 15 k $\Omega$		3)

1) Potential isolated from ground

2) Cover cut-out

3) The control voltage is measured at points 13, 14 and 15 after which resistor R<sub>p</sub> is placed in parallel. The voltage should not change if the frequency is locked



**Erläuterung (Beispiel):**

- ⑭ = Stromlaufplan 14 = circuit diagram 14
- 243-M = Leiterplatte M = circuit board M
- Pkt. 6 = Anschlußpunkt 6 = connect. point 6
- Tp. a = Testpunkt a = test point a
- T 1207/C Kollektor von T 1207 = collector of T 1207

**Note e.g. :**





Angegebene Pegel in Stellung "Eichen lin" gegen Masse gemessen  
 Indicated levels apply to instrument in position "CAL. LIN." and measured with respect to ground.

Bild (Fig.) 46 PM-5 / BN 243 und BN 243/1

Pegelplan dB, Np  
 (Signal Level Diagram)

Relaisplan (Relay Diagram)		Nummern der erregten Relais (Operating Relay No.)													eingeschaltete Dämpfung (insertion loss)						
dB- und Np-Gerät (dB and Np version)		Vorteiler (input attenuator)								Nachteiler (output attenuator)				log.	Vorteiler (input atten.)		Nachteiler (output atten.)		gesamt (total)		
Tastensatz S 1801 (push-button-set)	S 1803		rt	gn	gn	rt	rt	rt	gn	rt	1509	1510	1511	1512	1605 1606	dB	Np	dB	Np	dB	Np
	dB	Np	107	108	109	110	111	112	113	114											
Messen lin. (measure lin.)	-110	-12	x	x	x		x		x		x	x	x	x		-	-	-	-	-	-
			o	o	o		o		o		o	o	o	o		-	-	-	-	-	-
obere Zeile: x "rauscharm"	-100	-11	x	x	x		x		x		x	x	o	x		10	1	10	1	10	1
			o	o	o		o		o	o	o	o	o	o		20	2	20	2	20	2
untere Zeile: o "klirram"	-90	-10	x	x	x		x		x		o	o	o	x		20	2	20	2	20	2
			o	o	o		o		o		o	o	o	o		30	3	30	3	30	3
(upper line: x "low noise")	-80	-9	x	x	x		x		x		o	o	o	x		30	3	30	3	30	3
			o	o	o		o		o		o	o	o	o		40	4	40	4	40	4
(lower line: o "low distortion")	-70	-8	x	x	x		x		x		o	o	o	o		40	4	40	4	40	4
			o	o	o		o		o	o	o	o	o	o		50	5	50	5	50	5
	-60	-7	x	x	x		x		x		o	o	o	o		50	5	50	5	50	5
			o	o	o		o		o		o	o	o	o		60	6	60	6	60	6
	-50	-6	x	x	x		x		x		o	o	o	o		60	6	60	6	60	6
			o	o	o		o		o	o	o	o	o	o		70	7	70	7	70	7
	-40	-5	x	x	x		x		x		o	o	o	o		10	1	60	6	70	7
			o	o	o		o		o		o	o	o	o		70	7	60	6	80	8
	-30	-4	x	x	x		x		x		o	o	o	o		20	2	60	6	80	8
			o	o	o		o		o		o	o	o	o		70	7	10	1		
	-20	3	x	x	x		x		x		o	o	o	o		30	3	60	6	90	9
			o	o	o		o		o		o	o	o	o		70	7	20	2		
	-10	-2					x		x		o	o	o	o		40	4	60	6	100	10
							o		o		o	o	o	o		70	7	30	3		
	0	-1					x		x		o	o	o	o		50	5	60	6	110	11
							o		o		o	o	o	o		70	7	40	4		
	+10	0					x		x		o	o	o	o		60	6	60	6	120	12
							o		o		o	o	o	o		70	7	50	5		
	+20	+1					x		x		o	o	o	o		70	7	60	6	130	13
							o		o		o	o	o	o		70	7	60	6		
	-	+2					x		x		o	o	o	o		7		7		14	
							o		o		o	o	o	o		7		7			
Messen log. (measure log.) Kein Unterschied zwischen "rauscharm" und "klirram" (no difference between "low noise" and "low distortion")	-	+2					x		x		o	o	o	o	x	6					
							o		o		o	o	o	o	x	50	5				
	+20	+1					x		x		o	o	o	o	x	40	4				
							o		o		o	o	o	o	x	30	3				
	0	-1	x	x	x		x		x		o	o	o	o	x	20	2				
	-10	-2	x	x	x		x		x		o	o	o	o	x	10	1				
			o	o	o		o		o		o	o	o	o	x	-	-				
Eich.(cal.) lin.	-30	-4	x	x	x		x		x		o	o	o	o	x	-	-				
			o	o	o		o		o		o	o	o	o	x	30	3	60	6	90	9
Eich.(cal.) log.					x		x		x		o	o	o	o	x	30	3				

Bandbreite und Demodulationsart  
(Bandwidth and mode of demodulation)

Schalter- bzw. Tastenstellung (switch or push-button-position)	Relaisnummer (relay No.)				
	Umsetzoszillator (carrier oscillator)				
S 1701	S 1801/6	1704	1705	1706	1405, 1406
	3,5 kHz 500 Hz	x	x		x
	3,5 kHz 500 Hz			x	x
	3,5 kHz 500 Hz				x

Bei Umschaltung auf Leistungspegel kommt Relais 1513 hinzu.  
(Relay 1513 is also included in the power level mode).

x bzw. o = Relais erregt  
x resp. o = relay operative

Die Arbeitsweise der Stabrelais ist durch Farbpunkte gekennzeichnet.

Es bedeutet:

rot (rt) = Arbeitskontakt  
grün (gn) = Ruhekontakt

(The relay contacts are colour coded as follows:

red (rt) = normally open  
green (gn) = normally closed)

Reed-Kontakte im Abschlußwiderstand  
(reed switch for terminating impedance)

S 101 Z/Ω	Widerstand und Relaisnummer (resistance and relay No.)					
	150 Ω 101	300 Ω 102	487,5 Ω 103	75 Ω 104	1350 Ω 105	150 Ω 106
50	x			x		
60		x		x		
65			x	x		
75				x		
135					x	x
150						x

PM-5/BN 243, 243/1

Relaisplan (Relay Diagram)

## 5 CHECKING and ALIGNMENT

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### 5.1. Oscillator Portion RWO-5

#### 5.1.1. Instructions Regarding the Checking of the Power Supply

The transformer, switches and fuses are accessible on removing the unit from its cabinet. Therefore attention must be exercised to avoid electric shocks! !

Since the circuit board of the two point regulator 241-0 is mounted below the filter circuit board 241-R, it is necessary to release the latter by unscrewing and tilting it upwards.

The two chokes L 2405 and L 2406, the diodes GI 2402 and GI 2407, and the electrolytic filter capacitors C 2401 and C 2402 are accessible underneath the circuit board 241-0.

The connections of the power transformer are illustrated in section 5.1.1. "Mains Operation" of the Operating manual.

#### 5.1.2. Power Supply

Check the insulation resistance between the casing of T 2401 or T 2402 and the power supply chassis of the non-operating unit with the aid of an ohmmeter.

Nominal value:  $\infty$

This is to ensure that the two point regulator is "high" i.e. has no ground connection.

Checking the chopper:

Connect PS-5 and PM-5 as load for the power supply. Release circuit board 241-R by unscrewing and tilting upwards, thus making circuit board 241-0 accessible.

Disconnect the leads to the two point control at points 8 and 9 of circuit board 241-0 and feed the chopper from an external DC supply.

Set current limiting on the DC supply to 1 A.

Slowly increase the voltage at the power supply and observe whether the chopper commences oscillation.

A continuous tone should be heard during oscillation at input voltages from 4 V to approximately 8 V.



Measure the voltage to ground at point e of the additional filtering circuit board 241-AX.

Place the current limiting of the DC supply to 6 A and increase the chopper input voltage until 6.3 V is measured at point e, or 6.5 V for instruments of series A, B and C.

The input current should now amount to 5 A and the input voltage to 17.3 V.

Measure the voltages at the other connections of circuit board 241-AX against ground.

Point :	Nominal value :
c	+ 180 V
d	+ 37 V
f	- 2.9 V * (adjust with P 2501 of circuit board 241-AX)
g	- 19 V

Measure the chopper frequency between point 5 and point 1 of the circuit board 241-P with a frequency counter.

Nominal value: 28 kHz

\* For this alignment, the following adjustments are necessary at the PM-5:

Bandwidth to narrow, power level, demodulator position to level switch to - 110 dB (- 12 Np).



The voltmeter should have an accuracy of better than  $\pm 0,5$  %.

### 5.1.3. Checking of the two point regulator

Reconnect the chopper (point 8 and 9 of circuit board 241-0).

Rotate the maximum current control P 2401 and the voltage control to their fully clockwise positions.

Connect an ammeter with a measuring range of at least 6.6 A f.s.d. parallel to C 2404.

Switch on unit and adjust the short circuit current to 6.5 A.

Switch off measuring set and remove ammeter.

Connect the PSM-5 via a current meter to the AC line 220 V (120 V) and switch on. Adjust the voltage at point e of circuit board 241-AX to 6.3 V with potentiometer P 2402. Measure the AC line current.

Nominal value: 0.76 A (I .4 A)

- 5.1.6.      Checking the Output Signals of the Standard Frequency Buffer
- Connect an oscilloscope via a 10 : 1 probe to socket Bu 1102 of the buffer stage (11) (circuit board 241-AV).
- Oscillogram 2 should be displayed.
- Oscillogram 4 should be obtained at point 3.
- Oscillogram 5 should be available at point 2.
- 5.1.7.      Checking the Divided Frequencies
- Connect an oscilloscope via a 10 : 1 probe to point 1 of the 2.7 MHz crystal oscillator (17) (circuit board 241-K). Oscillogram 6 should be displayed.
- Measure at point 11 of the phase meter (14) (circuit board 241-AR).
- Oscillogram 7 should be displayed.
- Oscillogram 10 should be displayed at point 1 of the tuning indicator (19) (circuit board 241-AA) and oscillogram 3 observed at socket Bu 1101 of the standard frequency buffer stage (on circuit board 241-AW).
- 5.1.8.      Checking the Frequency of the 2.7 MHz Oscillator
- Disconnect the input of the mixer at point 3. Connect an oscilloscope via a 10 : 1 probe to point 2 of the frequency converter (20) .
- Oscillogram 14 should be displayed.
- Fed the 1 MHz standard frequency from socket Bu 1303 to the horizontal input of the oscilloscope. The displayed figure should be stationary.
- Reconnect the mixer input at point 3.
- 5.1.9.      Checking the Frequency of the Interpolation Oscillator, the 10 : 1 Frequency Divider and that of the Frequency Converter.
- Set the interpolation oscillator to 100 kHz.
- Place switch S 1602 to position 1.
- Connect an oscilloscope via a 10 : 1 probe to point 4 of the buffer and switching stage (21) (circuit board 241-H).
- Oscillogram 15 should be displayed.

measurement of the pull-in range. Select a vertical sensitivity of  $100 \frac{\text{mV}}{\text{cm}}$  and DC input.

Connect point 6 of the phasemeter to the oscilloscope ground connection and point 8 via a normal, unscreened measuring cable to the vertical input socket.

Lock the coarse frequency oscillator.

Tune slowly from the locked position and observe that the voltage displayed at the oscillograph will begin to increase. This voltage will return to 0 on leaving the pull-in range. The pull-in itself should occur at a voltage of at least 100 mV.

The pull-in voltage range should be symmetrical, i.e. the range above and below the locked position should be virtually equal.

Nominal value:  $\pm 100 \text{ mV}$ .

#### 5.1.12. Checking the Output Amplitude of the Auxiliary Oscillator

Tune coarse frequency oscillator to 0 MHz.

Connect an oscilloscope via a 10 : 1 probe to point 4 of the auxiliary oscillator discriminator (6) (circuit board 241-AD). Oscillogram 17 should be displayed.

#### 5.1.13. Adjustment and Checking of the Control Frequency Amplitude

Remove the plug connection at the control frequency output. Connect the RF millivoltmeter to point 5 or 6 of the control frequency buffer. Adjust the amplitude to 1 V with P 101.

Tune the coarse frequency oscillator to 5 MHz.

The coarse frequency is now tuned through the range up to 36 MHz, during which the amplitude should not deteriorate by more than 50 mV from the value obtained at 5 MHz.

Note: The control frequency supply at socket Bu 104 should not be allowed to operate under non-signal conditions if the plug connections at the output of the control frequency buffer. This is because a non-load condition would, due to the transformation via L 103, cause a considerable loading of the control frequency buffer, which would result in a reduction of the control frequency amplitude.

Tune and lock the coarse frequency oscillator to 16 MHz.

Align the frequency indicated on the counter with linearizing potentiometer P 703 to  $3 \text{ MHz} \pm 0.05 \text{ MHz}$ .

Tune and lock the coarse frequency oscillator to 0 MHz.

Align the frequency indicated on the counter with linearizing potentiometer P 704 to  $3 \text{ MHz} \pm 0.05 \text{ MHz}$ .

Tune the coarse frequency oscillator from 0 MHz through to 36 MHz and observe the frequency indication. It should read  $3 \text{ MHz} \pm 0.5 \text{ MHz}$  throughout the whole range. Reconnect lead to point 6.

#### 5.1.16. Characteristic Curve Alignment of Counter Discriminator I

Remove the sweep current lead from point 4 (circuit board 241-AU) of the frequency and phase control (8).

Connect point 2 to point 7.

Place switch S 1602 to position 13.

Connect an oscilloscope direct to the connection point of R 801, R 802 and C 803 of the frequency and phase control.

Tune interpolation oscillator to 45 kHz and coarse frequency oscillator to 0 MHz.

Align the difference frequency to minimum with trimmer C 912 of counter discriminator I.

Place interpolation oscillator to scale marker -10, adjust through to scale marker 100 and measure the difference frequency.

Nominal value:  $\pm 20 \text{ kHz}$ .

Reconnect lead to point 4.

#### 5.1.17. Checking the Frequency and Phase Control

Place switch S 1602 to position 1.

Connect the vertical input of the oscilloscope to point 9 and the horizontal input to point 10 of the frequency and phase control with unscreened measuring leads.

Switch the horizontal deflection to "ext".

Tune the coarse and interpolation oscillator throughout the whole range.

A sharp stationary, but strongly distorted ellipse should be displayed, whose position is slightly shifted on tuning the oscillators.

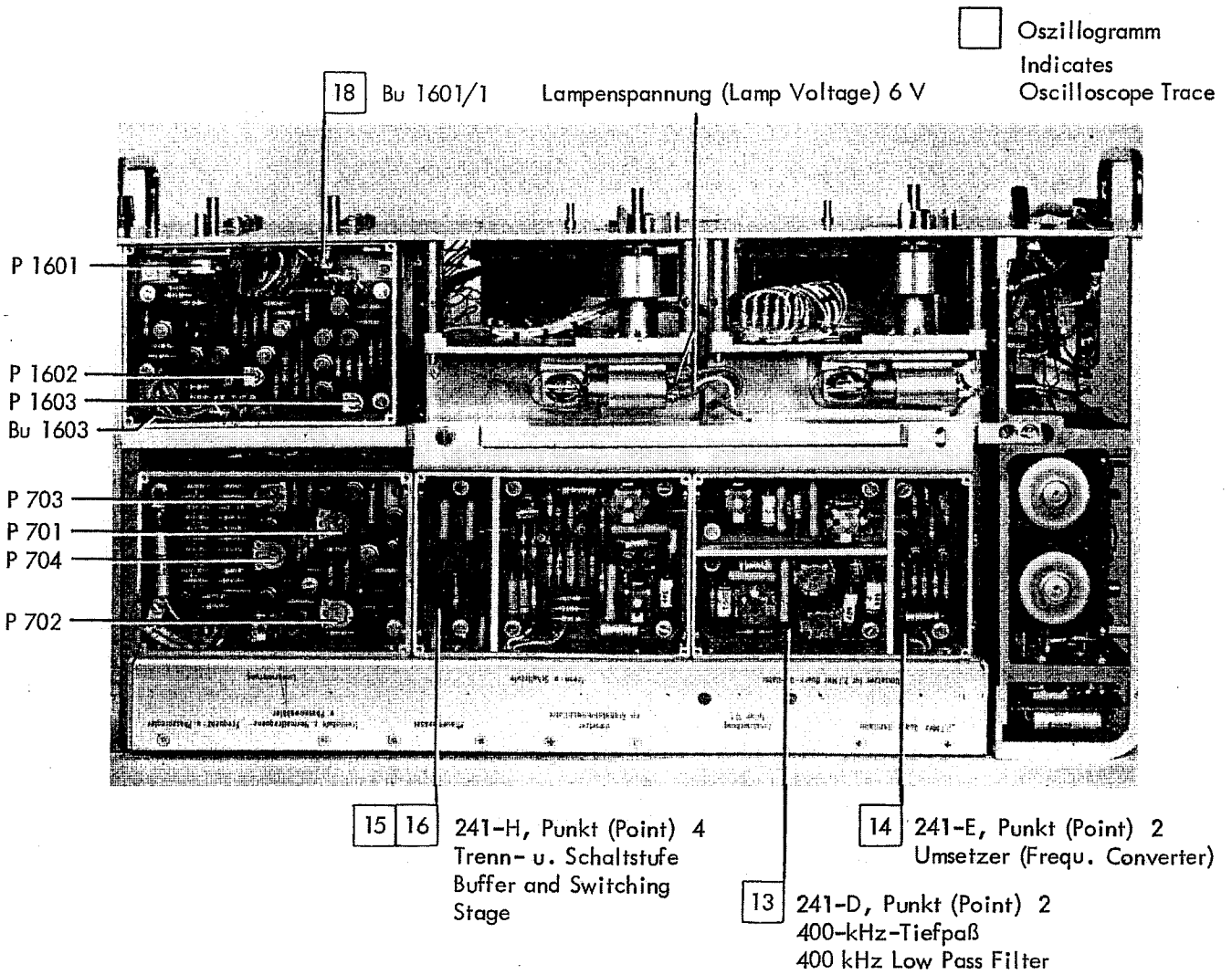


Bild 49 RWO-5 Ansicht von unten  
 Fig. 49 RWO-5 Bottom View

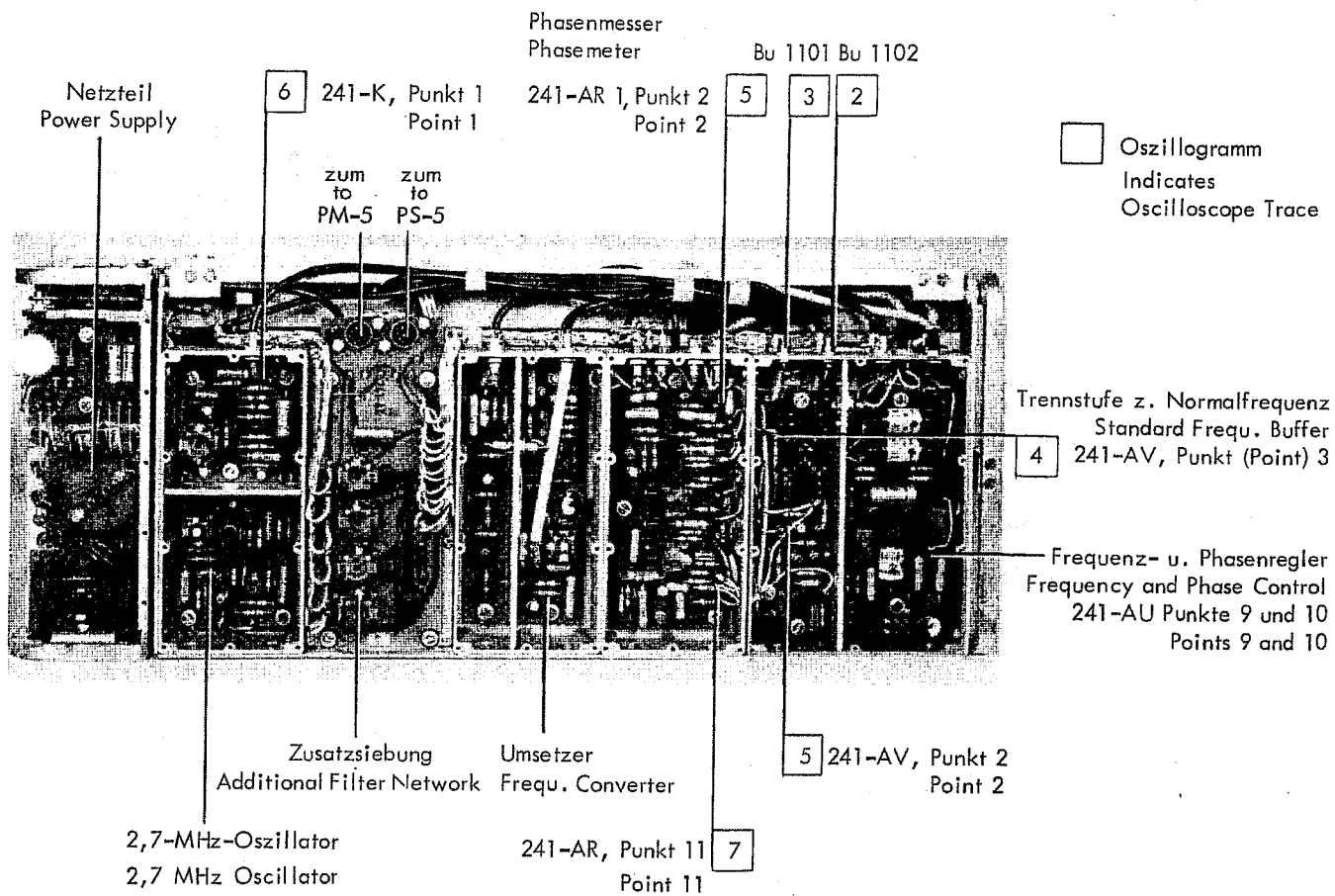


Bild 51 RWO-5 Rückansicht  
Fig. 51 RWO-5 Rear View

The aim of the alignment is to adjust P 201 and C 108 so that the hold range is approximately symmetrical to the control frequency and that the oscillator offers 69 MHz at a control frequency of 1 MHz. The cover of the oscillator sub-unit must remain closed during this alignment.

1. Disconnect the hunting oscillator and 69 MHz oscillator from point 5 of the phase control.
2. Place - 18 V on the disconnected lead to the 69 MHz oscillator. (This should cause a large detuning of the oscillator frequency by the varicap diodes).
3. Align the DC voltage at TP 202 and TP 203 with P 201 to read zero.
4. Reconnect the hunting oscillator and 69 MHz oscillator to point 5.
5. Adjust trimmer C 108 of the 69 MHz oscillator until the voltage at TP 202 and TP 203 is zero.

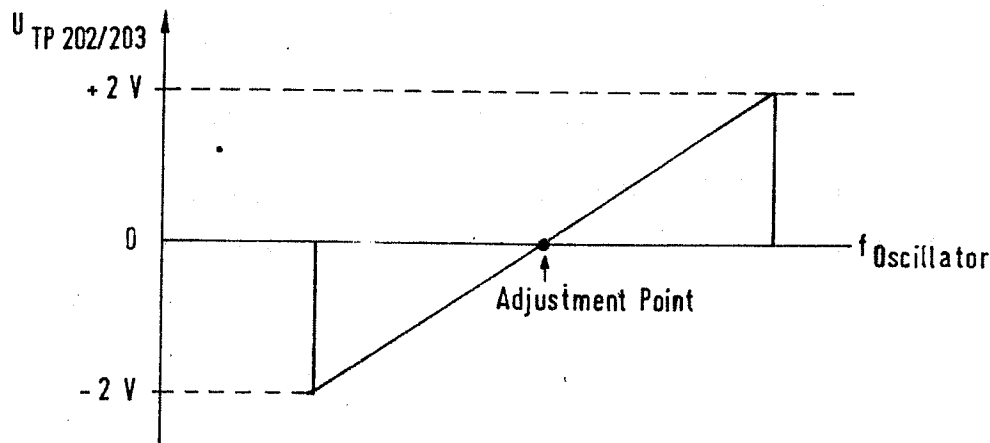


Fig. 53

Attention must be paid that this alignment is not carried out at 70 or 68 MHz. To establish this, set both frequency dials of the RWO-5 to 0 and ensure that the output signal of the PS-5 is absent. Lock trimmer C 108 with the locking nut.

### 5.2.3. Alignment of the 69 MHz Amplifier

Advance level potentiometer P 1002 until a voltage of approx. 20 mV is measured at test point 1 of the 69 MHz amplifier.

Measure the amplifier output voltage at test point 3 of the 69 MHz amplifier and align for maximum by alternately adjusting trimmers C 401 and C 406.

With pushbutton S 1801/7 depressed: The impedance value selected at S 101.

Accuracy of the DC values: 0.5 %

Resistor combinations:

Z / $\Omega$	Relay No.	Resistor R / $\Omega$
50	104, 101	75, 150
60	104, 102	75, 300
65	104, 103	75, 487,5
75	104	75
135	106, 105	150, 1350
150	106	150

The relay voltage must amount to at least 16 V. It is important that the threaded bolts are securely screwed into the pole shoes and that no air gap is formed which would increase the magnetic resistance.

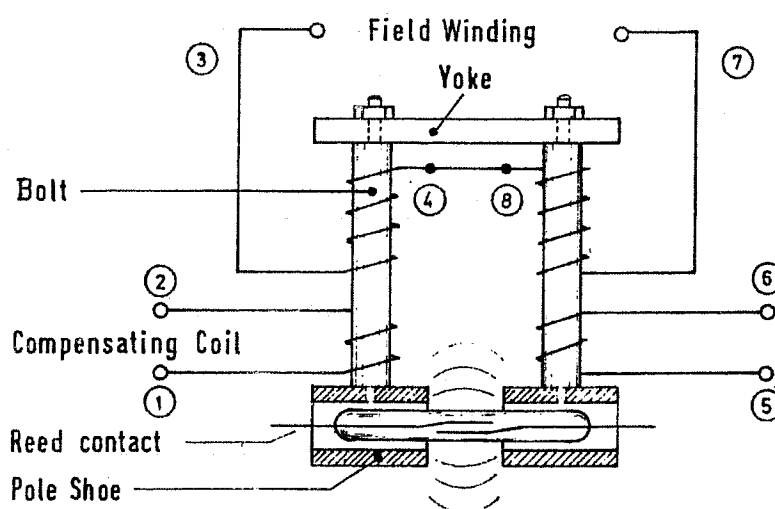


Fig. 54 Assembly of the magnetic portion

The reflection coefficient of the input resistors of an operational Level Meter PM-5 can be measured with a return loss measuring bridge (e.g. RFZ-5). An alignment is, however, only provided at 75  $\Omega$  impedance for a measuring frequency of 15 MHz. Large errors will also be seen as frequency response errors if the receiver is operated under terminated conditions.



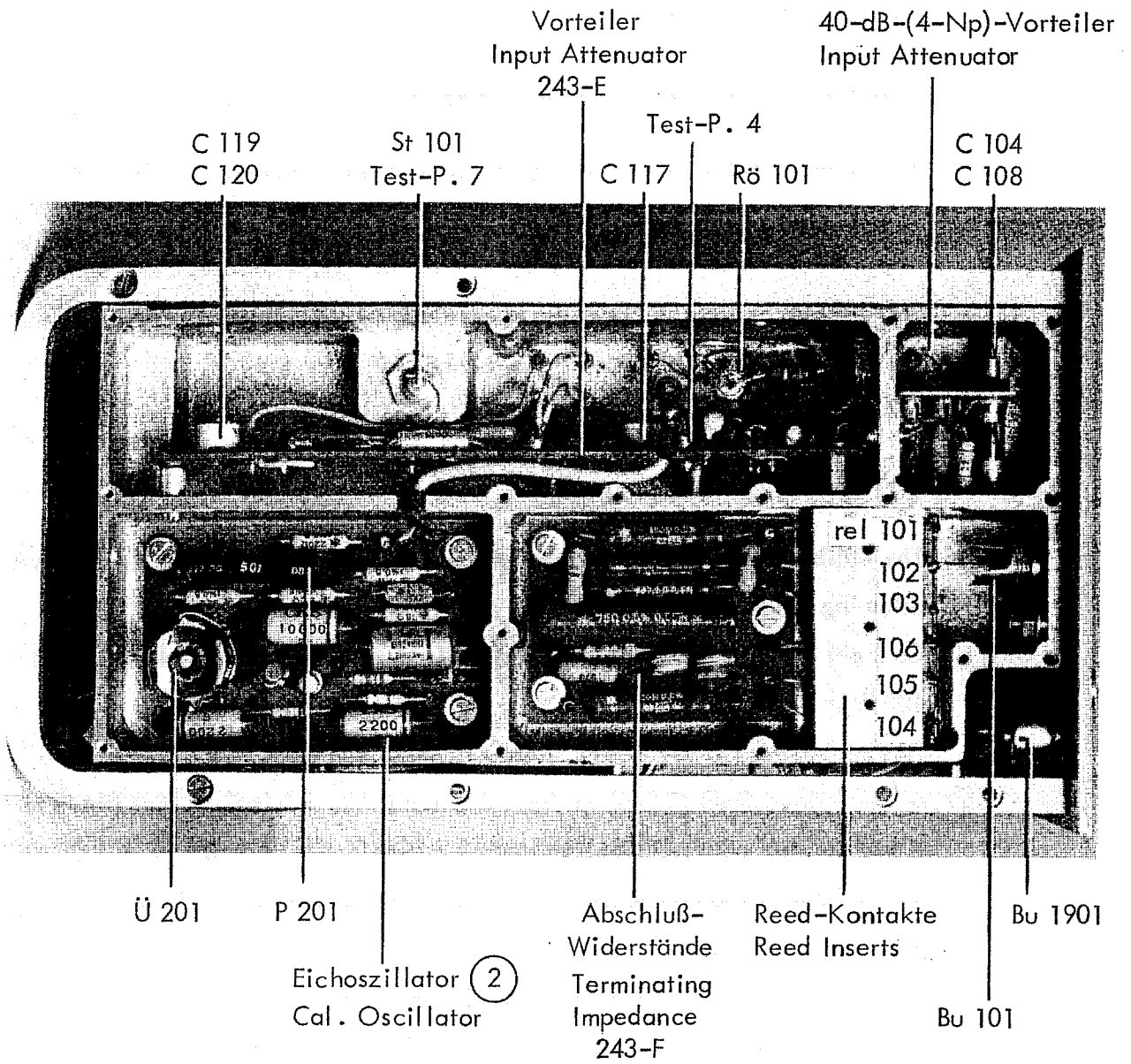
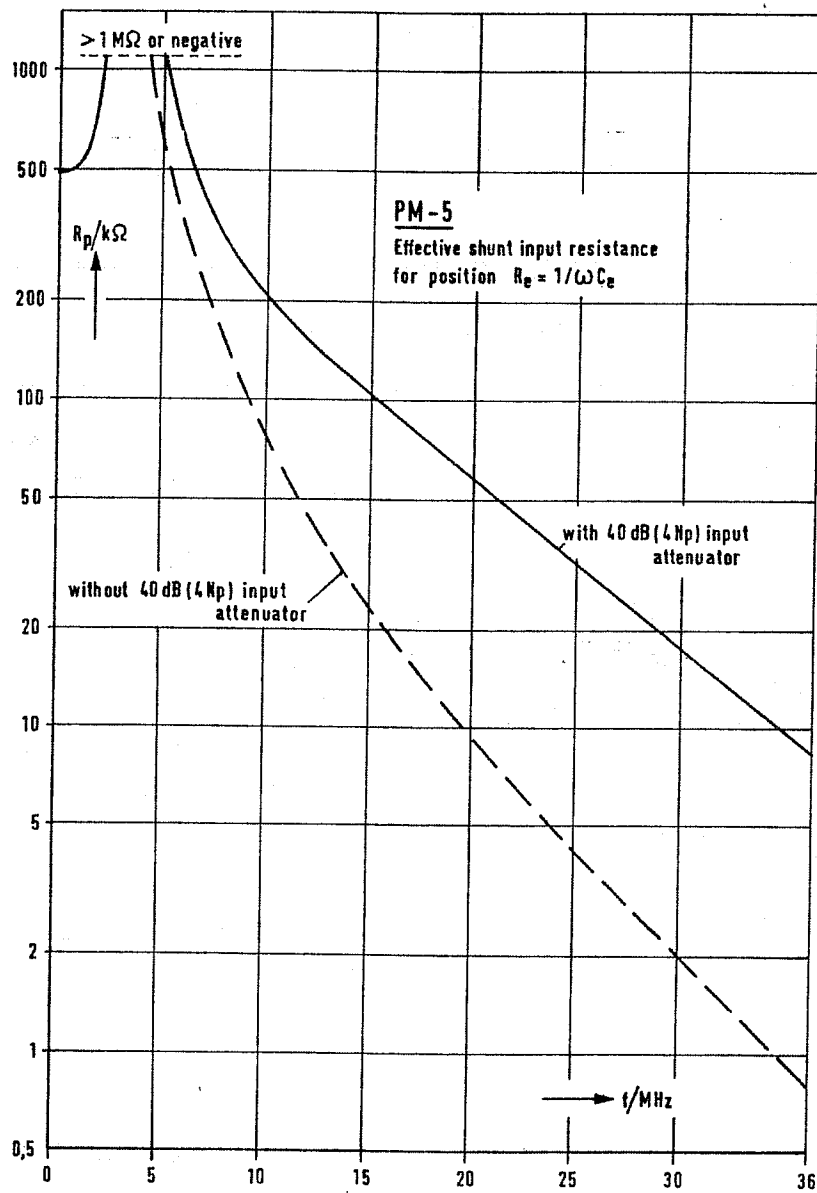


Bild 56 PM-5 Eingangsbaustein  
 Fig. 56 PM-5 Input Network



With 40 dB ( 4 Np ) input attenuator: + 20, + 10, ..., -10 dB ( + 2, + 1, ..., - 2 Np ) low noise  
+ 20, + 10, ..., -70 dB ( + 2, + 1, ..., - 8 Np ) low distortion

Fig. 58

Without 40 dB ( 4 Np ) input attenuator: all other sensitivity positions

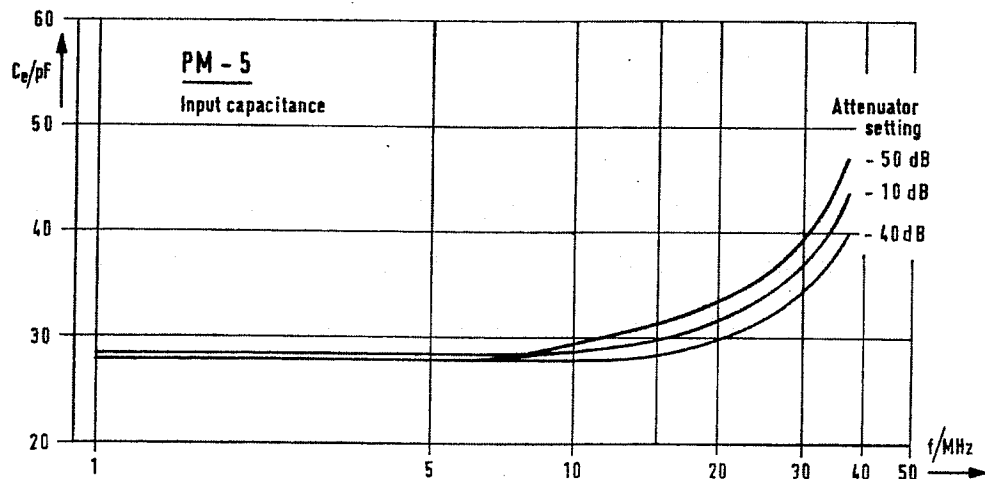


Fig. 59

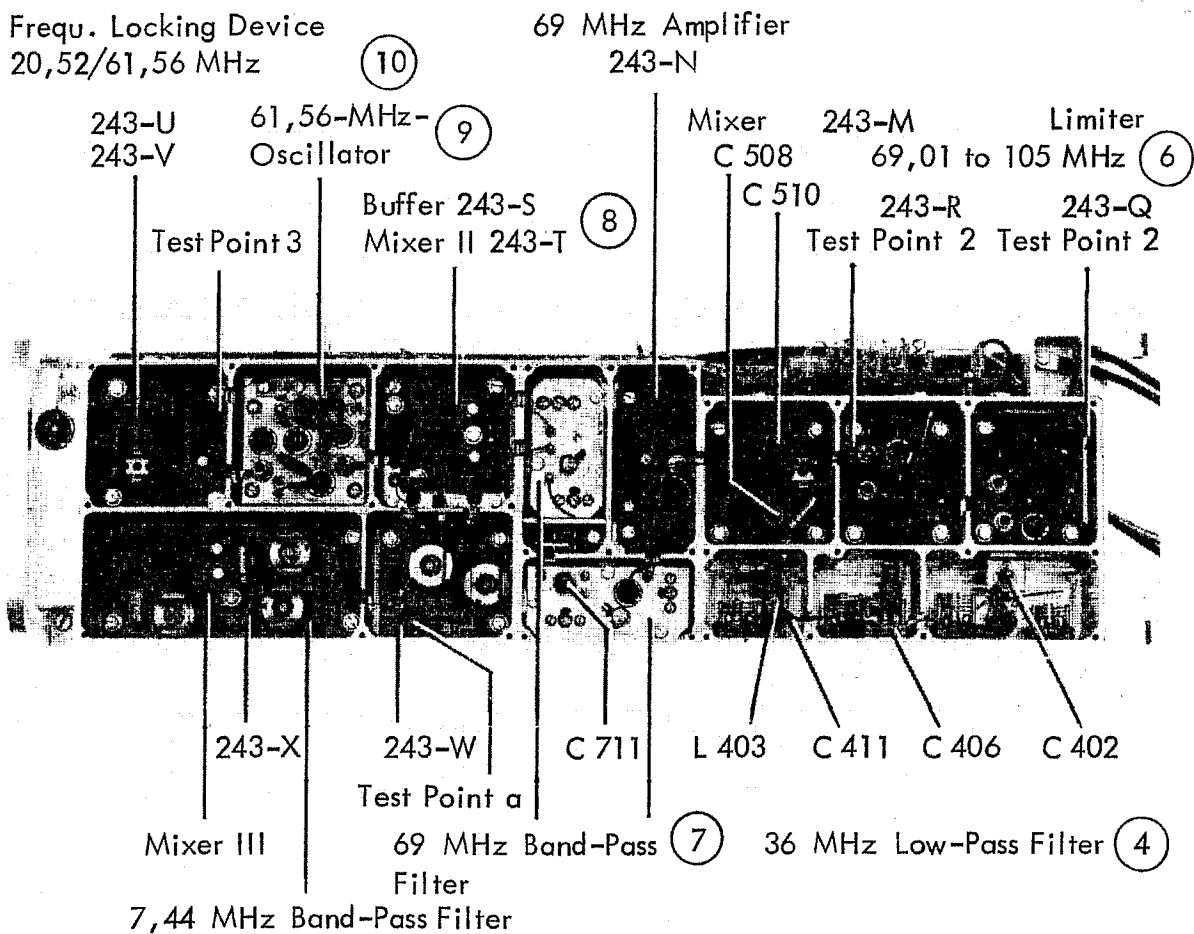


Fig. 60 PM-5 Rear View

a maximum reading at  $f = 613 \text{ kHz} \pm 50 \text{ Hz}$ . Repeat alignment and reconnect bridges.

### 5.3.10. 61.56 MHz Oscillator and Phase Locking Device

The third harmonic of the 6.84 MHz oscillator is fed via a bandpass filter to the frequency locking device. Connect an oscilloscope (eg. OS-1 with TK-4 from W u. G) to point 3 of circuit board 243-U and measure the 20.52 MHz voltage. Align Ü 1104 and Ü 1001 for maximum amplitude.

Nominal value:  $U_{pp} \approx 1.6 \text{ V}$

The oscillator voltages can be measured with the hp mV Meter type 411 A.

Nominal values Base of T 902: 0.8 V

Collector of T 902: 1.2 V

Collector of T 903: 1.0 V

For information regarding the alignment of the hold range see Section 4.3.2.2.

### 5.3.11. 6.84 MHz Oscillator and Phase Locking Device

The third harmonic of the 570 kHz crystal oscillator serves as synchronizing frequency. Connect an oscilloscope via the probe to point 5 of circuit board 243-O and align the 1.71 MHz voltage for maximum amplitude at Ü 1202.

Nominal value:  $U_{pp} \approx 3 \text{ V}$

For information regarding the alignment of the hold range see Section 4.3.2.2. and oscillograms 30 to 35 of Section 6.3.

### 5.3.12. 30 kHz Bandpass Filter

#### 5.3.12.1. Checking the selectivity curve of the PM-5, which essentially corresponds to that of the bandpass filter.

The easiest manner is to obtain the attenuation curve using the Sweep Measuring Set WM-50. Switch the PM-5 to "CAL": Calibration linear for checking the pass band, and logarithmic for checking the stop band range. Connect the SGE-10 to the PM-5 and RWO-5 and adjust the sweep width to the frequency range of interest. Centre frequency 100 kHz. The selectivity curves are given in the operating manual of the PSM-5. See also Section 1.1.3.4.

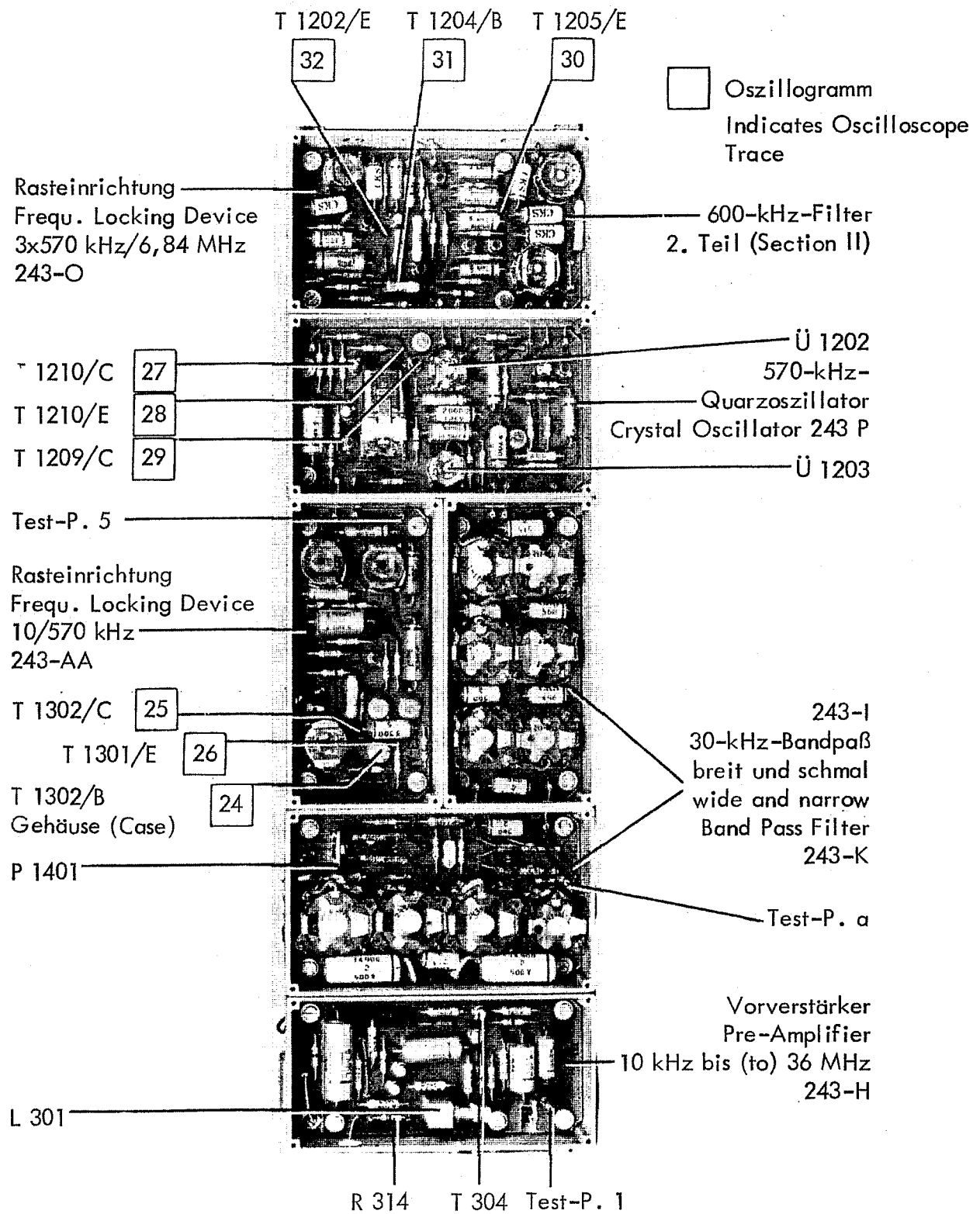


Bild 63 PM-5 Klappchassis Innenseite  
 Fig. 63 PM-5 Swing-Out Chassis (internal view)

range "0 dB (0 Np) log" and tune with the RWO-5 to 100 kHz. Align C 1602 for maximum reading.

Nominal value : - 100 dB (- 10 Np).

### 5.3.13.2. Logarithmic characteristic curve

Feed - 90 dB (- 9 Np) to the input of the PM-5. Adjust P 1601 (shift) to the scale marker - 90 dB (- 9 Np). Commencing with production series E, this adjustment is made with P 1903, whereby P 1601 should remain in its centre position.

Feed - 20 dB (- 2 Np) to the input and adjust P 1902 (expansion) to the - 20 dB scale marker.

Repeat alignment

During operation, the calibration is made only with potentiometer P 1601 to the calibration marker - 20 dB (- 2 Np) because the expansion of the characteristic curve remains unchanged. The range of adjustment should amount to  $\pm 8$  dB ( $\pm 1$  Np).

The linearity of the logarithmic scale can be checked with an attenuator test instrument or a calibrated attenuator.

The reference point is the - 20 dB (- 2 Np) scale marker. The measuring arrangement is as described in Section 4.3.3.1.

### 5.3.14. Output Sockets

The specified values given in Section 1.1.3.6. of the operating manual should be available at these sockets. A DC voltage is available between contacts 4 (-) and 5 (+) of sockets Bu 1601 and Bu 1603.

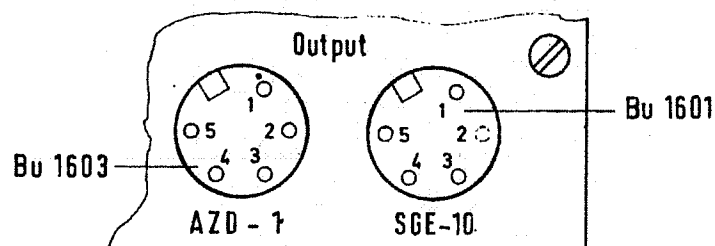




Fig. 65

### 5.3.15. Frequency Conversion Oscillator for the Demodulation of SSB Signals.

The oscillator frequency can be measured by connecting a frequency counter, via an isolating capacitor of 0.5  $\mu$ F, to point 1 of circuit board 243-A. A square wave

signal at an amplitude of  $U_{pp} = 220 \text{ mV}$  should be available. This voltage can also be measured at the slider of switch S 1701/III.

Frequency alignment:

Switching position		Frequency f/kHz	Align
S 1701	S 1801/6		
	3.5 kHz 500 Hz	28 29	Ü 1702 L 1701
	500 Hz 3.5 kHz	31 32	L 1702 L 1703

The bias current for the zero suppression amounts to approximately  $110 \mu\text{A}$  during logarithmic operation of the dB model and approximately  $160 \mu\text{A}$  for the Np model. Select the linear mode and connect a  $\mu\text{A}$  meter between point 3 and 4 of circuit board 243-A. The bias current can now be read from the  $\mu\text{A}$  meter.

#### 5.3.16. Voltage-free Front Panel

No galvanic connection should be present between the chassis and the front panel of the PM-5 if the power supply plug St 1901 and the control voltage are disconnected (remove plug connection at the RWO-5).

Connect an ohmmeter between the front panel and ground point of socket Bu 1602/1. Nominal value:  $\infty$

The RF isolation is made, during normal operation, by filter chokes in the ground connection and a choke in the control voltage channel.

### 5.4. Scale Expander AZD-1

#### 5.4.1. Opening the Cabinet

Remove the holding screws and detach rear cover.

The calibration potentiometer is to be found on the rear panel (Commencing with series D, this control is mounted on the front panel). It is thus necessary to carefully handle so that the connections are not broken.

#### 5.4.4. Checking the Attenuator

Select linear calibration at the PM-5 and adjust the calibration potentiometer for a reading of + 1 dB (+ 0.1 Np). Place the attenuation switch S 1 of the AZD-1 to + 1 dB (+ 0.1 Np). Adjust AZD-1 calibration potentiometer P 1 for a 0 reading. Reduce the reading on the PM-5 meter in steps of 1 dB with the calibration potentiometer P 1501, at the same time switching the AZD-1.

The reading on the AZD-1 must remain within  $\pm 0.06$  dB ( $\pm 0.006$  Np) of the zero marker.

#### 5.4.5. Checking the Expansion Adjustment (P 3)

Place the AZD-1 attenuation switch to - 2 dB (- 0.2 Np) and adjust the PM-5 calibration potentiometer for a - 1 dB reading. Adjust P 1 for a reading of +1 dB (+ 0.1 Np) on the AZD-1. Adjust the calibration potentiometer P 1501 for a -3 dB (- 0.3 Np) reading on the PM-5. If P 3 is correctly adjusted, the AZD-1 should now show a reading of exactly - 1 dB (- 0.1 Np). These tests can only be classed as a rough operational check and it is necessary to use special equipment, including a precision attenuator and a means of reconstructing the output level of the PM-5, to carry out an accurate calibration and check. This is especially true when the adjustment of P 2 for the balance of the differential amplifier or P 3 for the scale expansion is necessary. The following measuring arrangement can, however, be used with advantage:

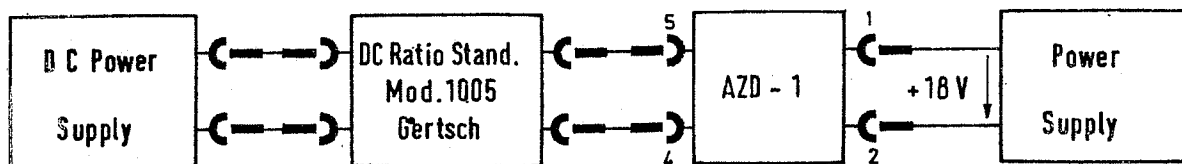


Fig. 67

Place AZD-1 potentiometer P 1 in its mechanical centre position. Place the attenuator switch to 0 dB (0 Np).

Adjust the DC power supply for a 0 reading on the AZD-1.

All divider positions can be checked with the aid of the following table:



#### 5.4.6. Alignment of the Attenuator

A warm-up period of at least 15 minutes is necessary before carrying out the alignment. Switch the attenuator selector of the AZD-1 to +1 dB (+ 0.1 Np).

Place potentiometer P 1 to its mechanical centre position. Set the DC ratio standard to 86 900 for the dB model or 90 500 for the Np model. Adjust the DC voltage until a 0 reading is obtained at the AZD-1. Set the attenuator selector to - 7 dB (- 0.7 Np) and adjust the DC ratio standard to 34 600 for the dB model or 40 700 for the Np model. Adjust the balance potentiometer P 2 for a zero reading at the AZD-1.

Check the zero adjustment in position + 1 dB (+ 0.1 Np) and realign if necessary.

Continue alternate alignment until the 0 point is correct in both positions. The alignment should always be ceased with P 2.

#### 5.4.7. Expansion Adjustment

Place potentiometer P 1 in its mechanical centre position.

AZD-1 attenuation switch to -2 dB (- 0.2 Np) and DC ratio standard to 69 000 (74 100).

Adjust the DC voltage for a reading of +1 dB (+ 0.1 Np) on the AZD-1. Set DC ratio standard to 54 800 (60 700) and adjust P 3 for a reading of -1 dB (- 0.1 Np) on the AZD-1.

Repeat alignment until the meter reading remains at full right and full left scale on switching the DC ratio standard.

### 5.5 Probe TK-8

#### 5.5.1. Current Consumption and Operating Points

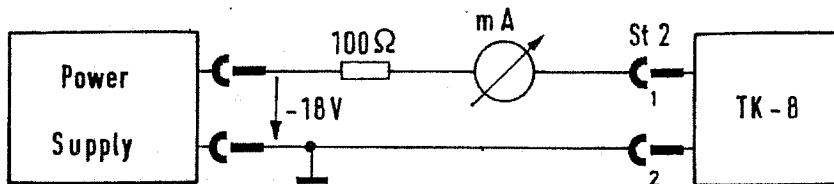


Fig. 68 Measuring arrangement

Connect the TK-8 via a dropper resistor to -18 V.

Measure the current consumption.

Nominal value: 50 mA.

The following additional equipment is required: A signal generator with a frequency range of 100 kHz to 36 MHz and a harmonic distortion factor better than 1 %, together with a RF millivoltmeter to monitor the input signal of the probe, which must remain constant over the whole frequency range.

The alignment of the probe is carried out in the 0 dB (- 1 Np) position of the PM-5.

Adjust P 1 and C 2 for the flattest frequency response curve (P 1 up to 15 MHz and C 2 at 36 MHz).

Nominal value: max. deviation from ideal curve  $\pm 2 \%$ .

2. This alignment is simplified if the TK-8 is to be used with a PSM-5, because the input level for the probe is automatically held at a constant level throughout the whole frequency range.

Set the PS-5 for a -10 dB (- 1.15 Np) output level reading. A level of - 4 dB (- 0.46 Np) is thus available at the generator output and the harmonic distortion factor remains less than 1 %.

3. The best method is however, to display the frequency response curve during swept frequency operation. For this, a Display Unit SG-1 and the matching Plug-in Unit SGE-10 are required in addition to the PSM-5.

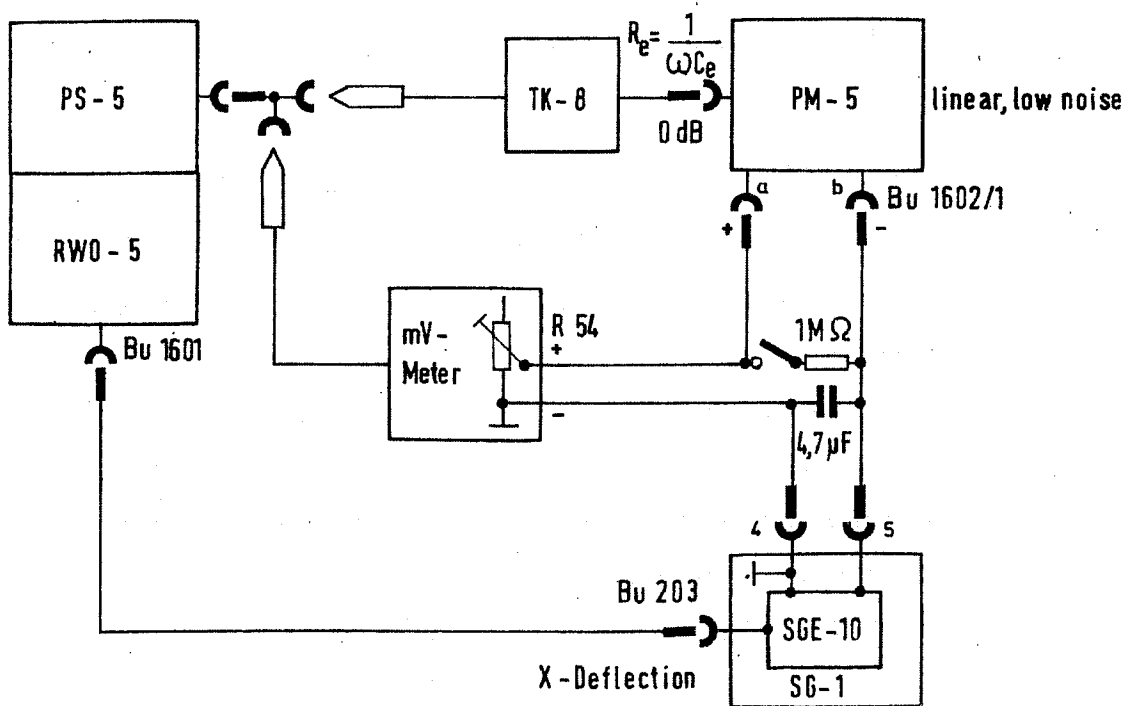


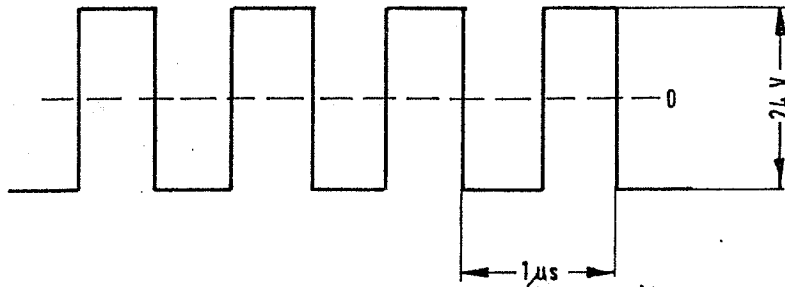
Fig. 70 Measuring arrangement

The attenuator position + 10 dB (0 Np) low noise corresponds the - 10 dB low distortion position.

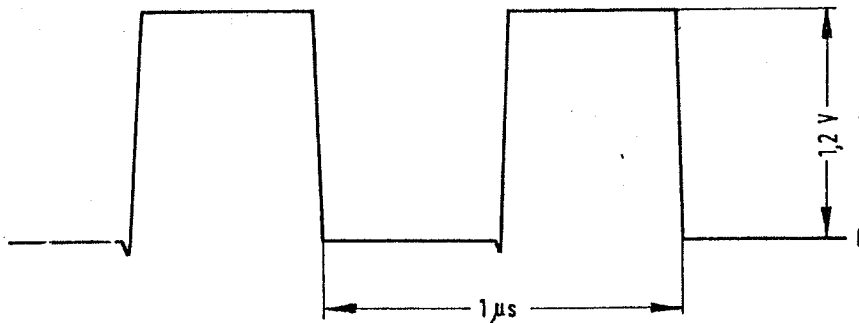
A voltage of - 3 V is fed to point 5 of circuit board 243-L for the + 10 dB and 0 dB (0 Np and - 1 Np) values.

Nominal value : The deviation from the 0 dB (- 1 Np) curve should be within  $\pm 1 \%$ .

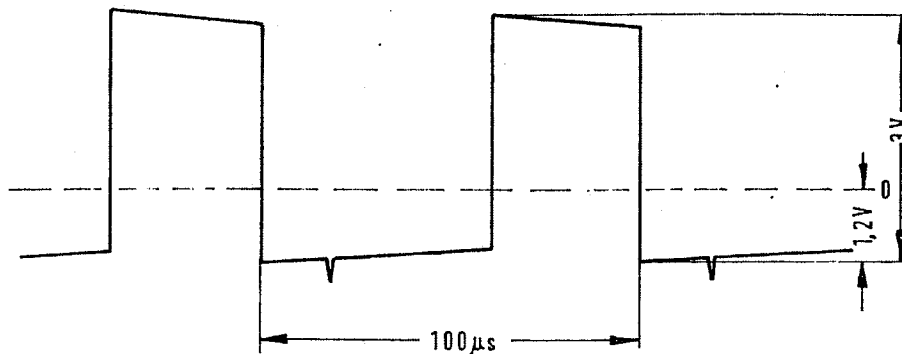
1 Normalfrequenz-Oszillator  
Standard Frequ. Oscillator (13) BU 1303



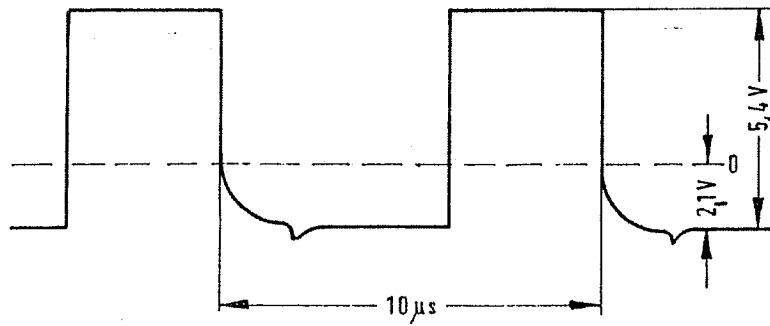
2 Trennstufe z. Normalfrequenz  
Standard Frequ. Buffer (11) BU 1102



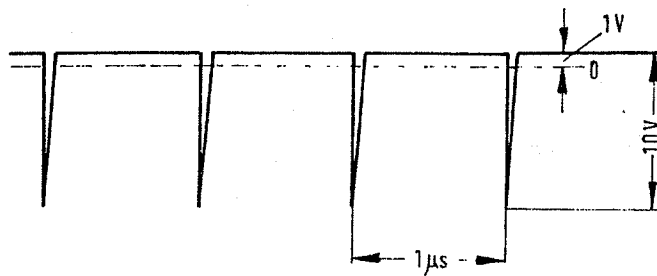
3 Trennstufe z. Normalfrequenz  
Standard Frequ. Buffer (11) BU 1101



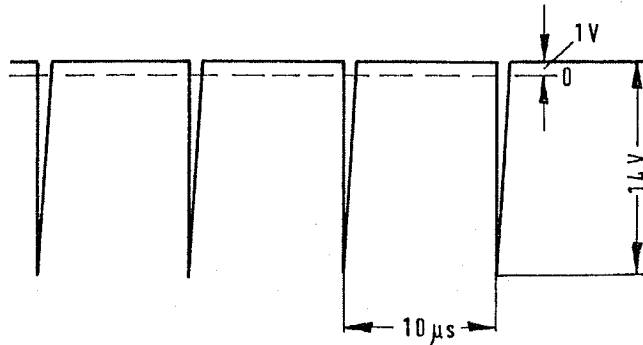
7 Phasenmesser (14) 241-AR Punkt (point) 11  
Phase Meter



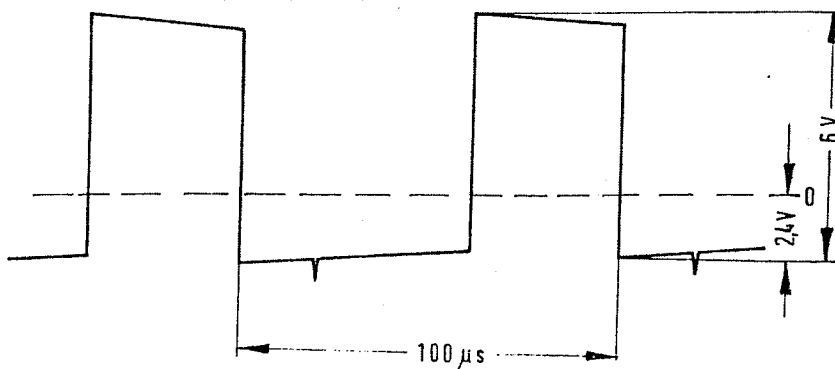
8 Phasenmesser (14) 241-AR Kollektor (collector) T 1403  
Phase Meter



9 Phasenmesser (14) 241-AR Kollektor (collector) T 1405  
Phase Meter

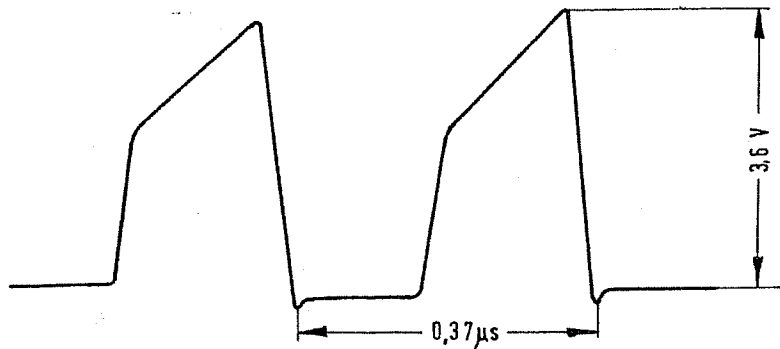


10 Abstimmanzeige (19) 241-AA Punkt (point) 1  
Tuning Indication



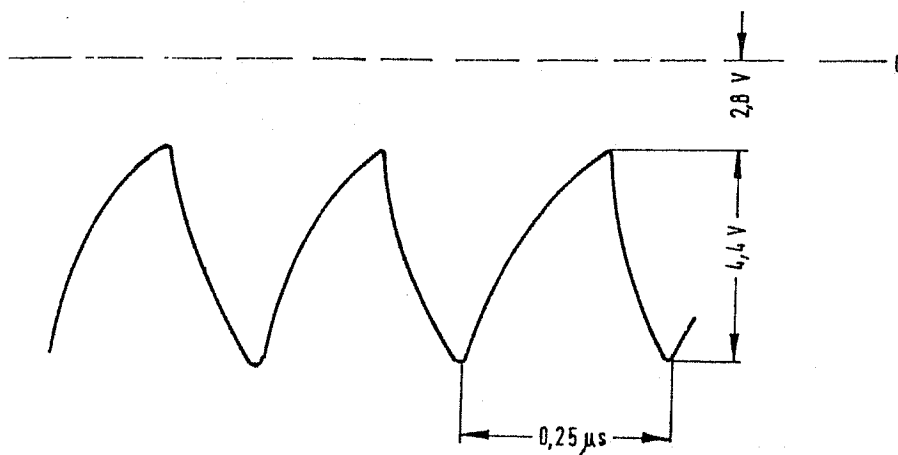
14 Umsetzer  
Frequ. Converter

20 241-E Punkt 2 (Punkt 3 abgelötet!)  
Point 2 (Point 3 disconnected)



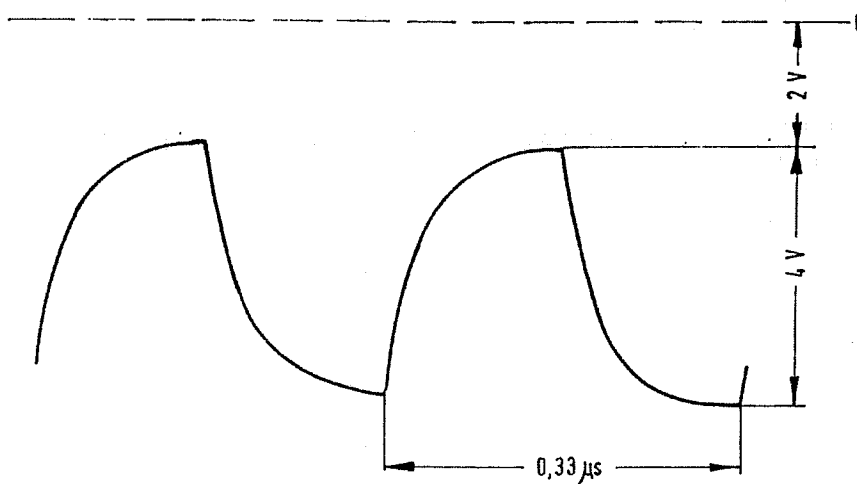
15 Trenn- u. Schaltstufe  
Buffer and Switching Stage 21 241-H Punkt (point) 4

Interpolationsoszillator auf 100 kHz eingestellt, S 1602 in Stellung 1  
Interpolation Oscillator set to 100 kHz, S 1602 in Position 1

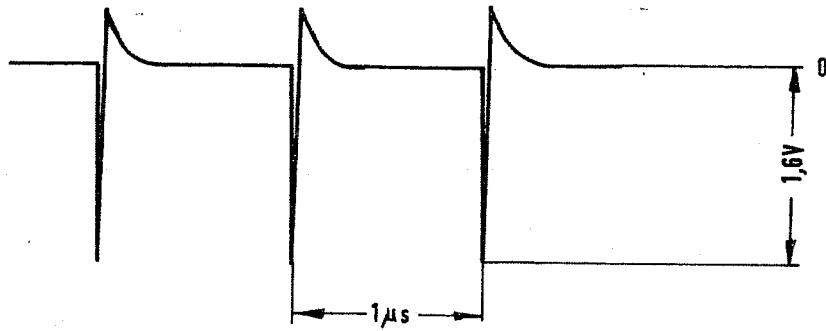


16 Trenn- u. Schaltstufe  
Buffer and Switching Stage 21 241-H Punkt (point) 4

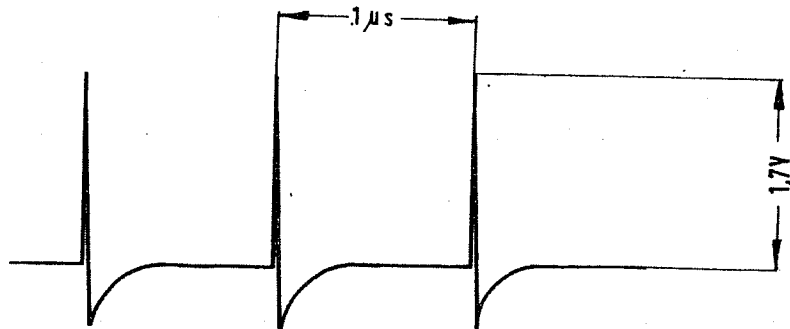
Interpolationsoszillator auf 100 kHz eingestellt, S 1602 in Stellung 2  
Interpolation Oscillator set to 100 kHz, S 1602 in Position 2



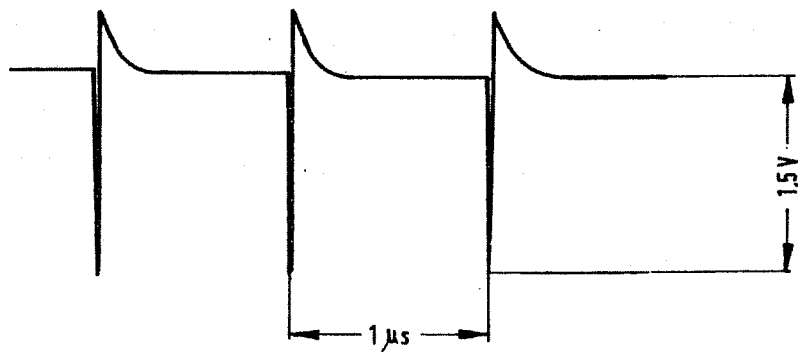
21 Phasenregler (2) 242-E Primärseite Ü 201 Anschluß 1  
Phase Control Primary Side Ü 201 Connection 1



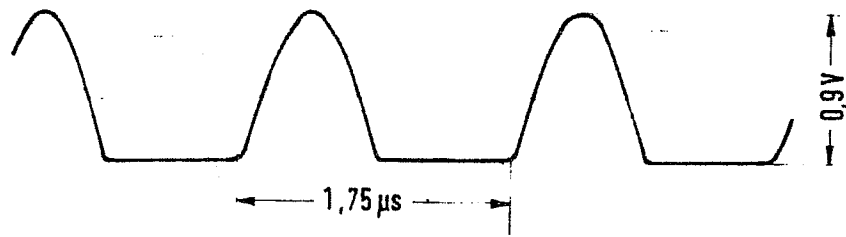
22 Phasenregler (2) 242-E Sekundärseite Ü 201 Anschluß 6  
Phase Control Secondary Side Ü 201 Connection 6



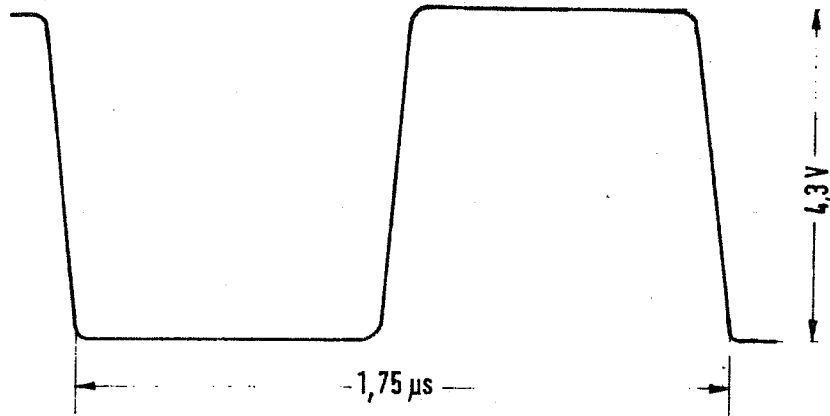
23 Phasenregler (2) 242-E Sekundärseite Ü 201 Anschluß 3  
Phase Control Secondary Side Ü 201 Connection 3



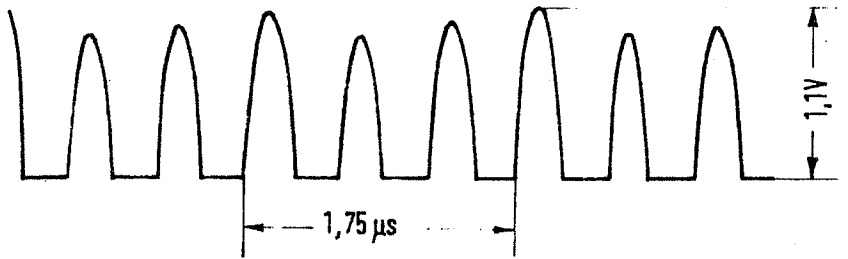
28 Umsetzung 600 kHz/30 kHz (12) 243-P 1 Emitter T 1210  
Frequ. Converter



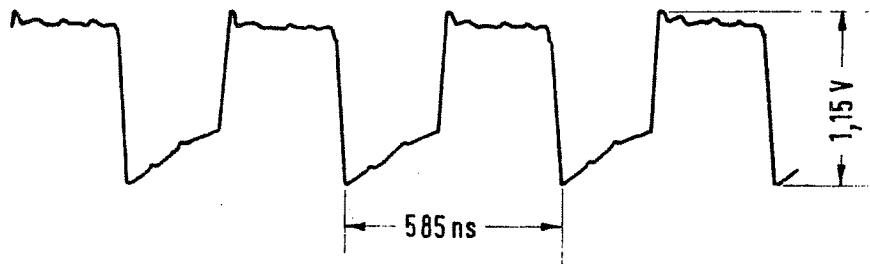
29 Umsetzung 600 kHz/30 kHz (12) 243-P 1 Kollektor T 1209  
Frequ. Converter Collector T 1209



30 Umsetzung 600 kHz/30 kHz (12) 243-0 Emitter T 1205  
Frequ. Converter



31 Umsetzung 600 kHz/30 kHz (12) 243-0 Basis T 1204  
Frequ. Converter Base T 1204





## 7. ADJUSTMENT OF THE OPTICAL PORTION AND THE DRIVE OF THE OSCILLATOR PORTION

These adjustments should normally not be carried out by the customer. If it should, however, be impossible to dispatch the instrument to the factory or a qualified service depot, the adjustment and repair can be carried out according to the following instructions.

### 7.1. Condenser and Lens

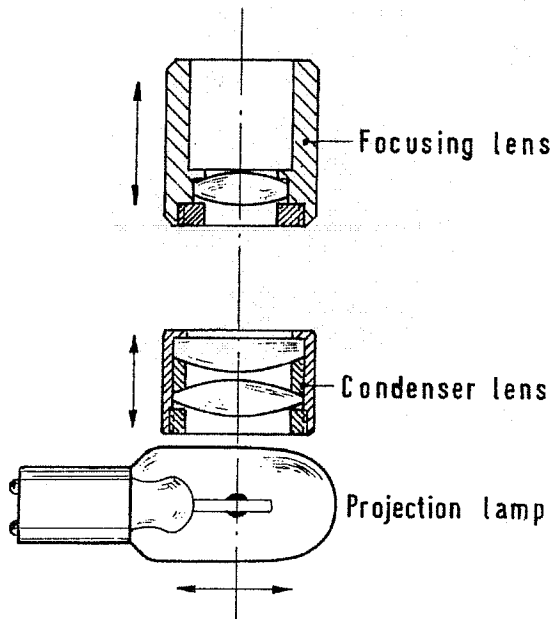


Fig. 71

The lens, condenser and projector lamp are assembled in the drive system as shown in Fig. 71. The focus and the illumination of the transparent scale can be adjusted by shifting the lens and condenser in an axial direction. The die-cast casing of the drive system is provided with two cut-outs, through which the holding screws of the lens and condenser are accessible. If these allen screws are loosened with the corresponding allen key (metric SW 1.5), the lens or condenser can be adjusted with the adjustment tool.

The adjustment tool is knurled under  $45^\circ$  which moves the lens holder when rotated in the cut-out adjacent to the holding screw.

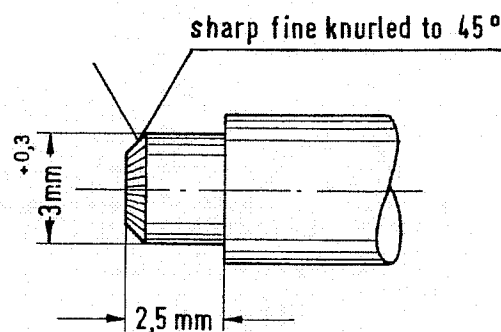


Fig. 72

The focussing is carried out by adjusting the lens position.

An unequal illumination on the transparent screen can be avoided by adjusting the condenser. If the unequal illumination cannot be cured with the condenser itself, it will be necessary to adjust the projection lamp.

This is described in Section 8.3.

## 7.2. Mirror Adjustment

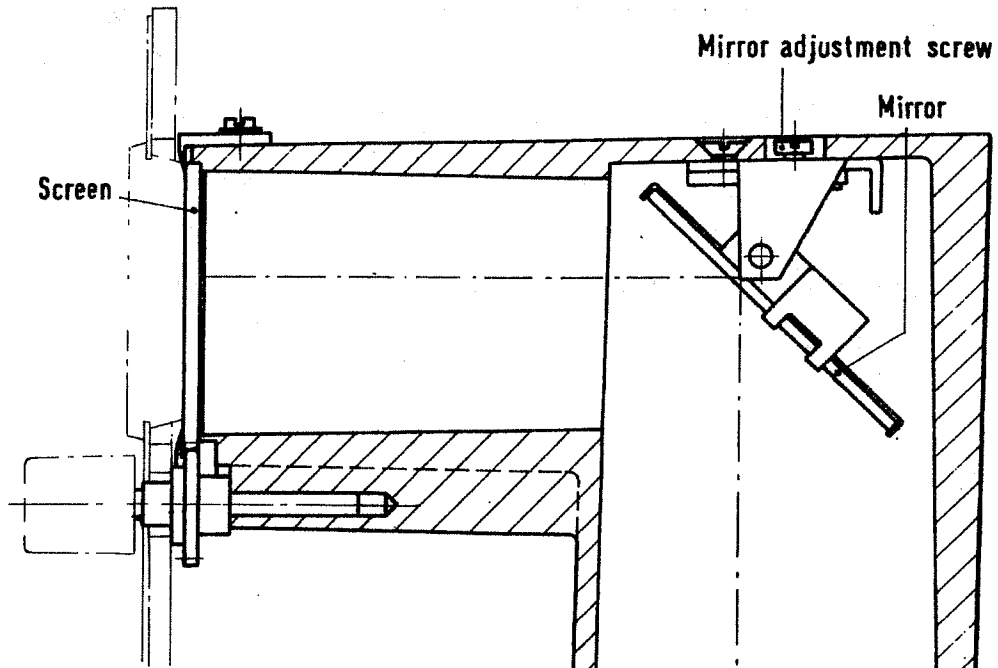


Fig. 74

The mirror can be adjusted in the horizontal and vertical planes. The vertical adjustment is made with the mirror adjustment screw, thus adjusting the length of the projected scale.

It will be necessary to adjust the mirror in horizontal plan if the scale is not truly vertical. To do this, the three Philips screws on the upper side of the casing are loosened and the mirror is adjusted by means of a wooden rod, which has been inserted into the shaft. The 3 screws should be retightened after the adjustment.

### 3. INSTRUCTIONS REGARDING MAINTENANCE AND SERVICE

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#### 8.1. Switch Care

The switch axles of the rotary switches should be lightly oiled if a stiffness is observed at a later date.

It is also recommended to clean the switch contacts roughly once a year with a chamois cloth.

If the contacts are soiled, they should be cleaned with alcohol.

#### 8.2. Cleaning of the Lenses

If a brightness reduction of the projection scale is noticed, the condenser lenses should be checked (located directly above the projection lamp).

It is firstly necessary to remove the projection lamp from its holder, which is carried out, with the bayonet type, by depressing the glass bulb and simultaneously rotating to the left. For instruments equipped with the clamp holder, the projection lamp is simply removed by lifting the clamp spring with a screwdriver or coin. The most favourable cloth for cleaning the lens is a chamois material.

#### 8.3. Exchange of the Projection Lamp

If it is necessary to exchange a defective or insufficiently bright projection lamp, this is carried out as described in the previous section.

A replacement lamp is provided with the instrument and is to be found in a bracket on the casing of the sawtooth oscillator. If an unequal illumination is observed on the transparent screen, it will be necessary to adjust the position of the lamp. The lamp holder should be adjusted so that the filament directly coincides with the condenser axis.

This can be easily seen by observing the transparent screen.

The filament must also be parallel to the surface of the lens, otherwise a equal illumination of the screen will not be achieved. This is achieved by rotating the lamp holder in its bracket. The glass bulb of the lamp should be cleaned with a cloth dampened in alcohol.