

THE EFFECT OF TRANSFORMERS ON TRANSMISSION OF DIGITAL AUDIO SIGNALS

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Transformers used in the transmission of digital audio signals affect signal fidelity, interference susceptibility, and conducted EMI emission. Commercially-available transformers specified for digital audio exhibit vast differences in performance. Transmission standards often specify only transformer ratio and bandwidth, and ignore many other parameters which affect critical applications. This paper reviews the function, parameters and performance of digital audio transformers, and presents data on frequency response, pulse aberration, common mode rejection ratio, and jitter. It also discusses several applications and compares a number of transformers in detail.

1. Transformers in Digital Audio Transmission Systems

Digital audio transmission systems use transformer coupling to provide balanced outputs, improve common mode noise rejection, match impedances, and reduce conducted EMI emission and susceptibility. (Note: Throughout this paper, the term “digital audio signal” refers to the Manchester encoded type of modulation with embedded clock, etc. as specified in AES/EBU, SPDIF, AES-3 and the corresponding twice speed signals such as DVD, etc.)

Figure 1 is a block diagram of a direct-coupled transmission system, without transformers. The signal fidelity is affected by transmitter slew rate, the cable, and the pickup of common-mode interference along the cable. The receiver’s differential input sees the common-mode noise appearing between the transmitter and receiver grounds. The performance of the receiver circuit depends on the levels of the common-mode interference and the signal. Interference sources include high speed DSP and microprocessor clocks, RF noise, switching power supplies, and crosstalk from adjacent cables.

The receiver IC’s differential amplifier is characterized by the common-mode rejection ratio (CMRR) which decreases with increasing frequency. Because high frequency noise is capacitively coupled to the cable (crosstalk), a direct-coupled input is highly susceptible to such noise. Another characteristic of direct coupling is that EMI and crosstalk on the cable shield can enter and contaminate the transmitter’s or receiver’s internal circuitry (e.g. power, clocks or ground planes) if the connector shield or cable return is grounded to the circuit ground.

Figure 2 is a block diagram of a transmission system using transformers. The transmitter’s digital signal is coupled to the output through a transformer. (Note that the figure shows a balanced cable but single-ended coax can also be used.) The transformer output is isolated from the chassis ground. The output is connected to a balanced cable, and the cable is connected to a transformer-coupled receiver. The resistors set the source and termination impedance to match the characteristic impedance of the cable.

The insertion of the transformer in the receive circuit greatly improves the high frequency CMRR, thus reducing recovered clock jitter. Even receivers with high jitter attenuation will benefit, since the transformer also attenuates interference contamination by common-mode noise. Both transmitter and receiver sides make use of transformer coupling to break ground loops, reduce conducted EMI, and provide voltage or impedance matching.

2. AES/EBU Signal Bandwidth

Frequency components of the digital audio signal exist far beyond the minimum 100 kHz - 8 MHz bandwidth often specified [1, 2, 3]. Figure 3 is a spectrum of an AES/EBU signal, with 48 kHz sample rate (F_s), swept from 0 to 100 kHz with vertical scale of 10 dB/div. Note the spectral content below 100 kHz. Figure 4 is a spectrum of the same signal from 0 to 50 MHz, displaying substantial energy above 8 MHz.

These spectra show that the digital audio signal bandwidth extends far beyond the minimum bandwidth usually specified. Although the minimum bandwidth provides useable transmission, recovery of low-jitter clocks and accurate data transmission in real-world noisy environments will benefit from greatly increased bandwidth. The transformer is often the limiting factor in the end-to-end bandwidth of such a system.

An extension beyond the minimum by 5 to 10 times for both upper and lower bandwidths, (e.g. from 5 to 20 kHz F_{LOW} and 50 to 100 MHz F_{HIGH}) is recommended. For twice speed applications, for example, 96 kHz and DVD, F_{HIGH} may be increased to 200 MHz.

3. Transformer Equivalent Circuit

Figure 5 is a generalized equivalent circuit for any transformer [4]. All components except the ideal 1:N transformer represent parasitic effects. R_P and R_S are the resistance of primary and secondary windings. L_{PLKG} and L_{SLKG} are the primary and secondary leakage inductances caused by the separation in space of the two windings. Capacitance C_{PSHUNT} and C_{SSHUNT} are the primary and secondary inter-winding shunt capacitance. Inductance L_{PMAG} is the primary magnetizing inductance, representing the self-inductance of the primary winding when no secondary current flows. R_{PLOSS} is the core loss component representing energy dissipated by the core as it is magnetized and de-magnetized. The inter-winding capacitance from primary to secondary is C_{P-S} .

These parasitic components, in combination with the ideal transformer and the impedance of the external circuit environment, form a broad bandpass filter with typical F_{HIGH} / F_{LOW} ratios of 1000 to 10,000. The low-frequency corner, F_{LOW} is determined by the R - L circuit formed by the source impedance plus R_P in combination with the magnetizing inductance L_{PMAG} . The high-frequency corner F_{HIGH} is defined by the RLC circuit including shunt capacitances C_{PSHUNT} and C_{SSHUNT} , series leakage inductances L_{PLKG} and L_{SLKG} and the circuit environment's impedance. This lumped constant model is a second-order approximation and does not account for pulse aberrations due to the distributed nature of the windings.

Figure 6 is a computer simulation of the frequency response of the model in a 50 Ω environment. The primary inductance, L_{PMAG} is varied from 200 to 1400 μH , (typical values) causing the low frequency 3-dB point F_{LOW} to be reduced from 20 kHz to 3.3 kHz. Figure 7 is a similar plot where the series leakage inductance L_{PLKG} changes from 0 to 300 nH. The high-frequency corner, F_{HIGH} , is reduced from 160 MHz at 100 nH to 50 MHz at 300 nH.

For maximum bandwidth, we recommend choosing a transformer with the highest possible primary inductance L_{PMAG} and lowest possible leakage inductance, L_{PLKG} . Since leakage inductance is difficult to measure and specify, the high-frequency corner, F_{HIGH} can be used to specify that parameter.

4. Frequency Response and Saturation Effects

Figure 8 shows the effect of the transformer's high-frequency corner on the pulse risetime. The upper trace is a low aberration pulse generator. The middle trace is a transformer with F_{HIGH} equal to 100 MHz and the lower trace is a transformer with F_{HIGH} equal to 35 MHz. The benefit of wider bandwidth is obvious.

Figure 9 shows the effect of a transformer's low-frequency corner, F_{LOW} , on the eye pattern of a received digital audio signal. The upper trace was taken with a transformer with F_{LOW} of 4 kHz and the lower trace was taken with a transformer with an F_{LOW} of 400 kHz. Note the closing of the eye pattern in the lower trace, which causes an increase in inter-symbol interference. This highlights the importance of using transformers with an extended F_{LOW} .

Figure 10 illustrates the effect of saturation of the transformer core on the signal. This effect is nonlinear; it is not accounted for in the model above, but is a function of the voltage and time duration of each pulse. The maximum flux density is a property of the core material, and determines the point at which the transformer action ceases. The saturated core has a primary inductance, L_{PMAG} near zero, thus shorting out both primary and secondary.

The transformer in the upper trace has a large flux capacity of 300 μVs , and exhibits no saturation. The lower trace is a transformer with similar frequency response but only 50 μVs flux capacity, showing severe saturation. The flux capacity must be maximized to avoid such problems. Regardless of the transformer's flux capacity, a capacitor is required in series with the transformer's primary to prevent DC bias from causing saturation.

5. Pulse Aberration

Nonlinear phase vs. frequency response creates pulse aberration. This may be present even in transformers with relatively wideband, flat magnitude of frequency response. Figure 11 was obtained with a 12.288 MHz, 1.5 ns risetime pulse generator (12.288 MHz is the fastest symbol rate in a 96 kHz digital audio signal). The upper trace is a very linear phase response transformer with minimal aberration. The lower trace is a transformer with similar bandwidth but with substantial phase nonlinearity; note the severe overshoot and pulse aberration.

6. Common-Mode Rejection Ratio and Interference Suppression

Figure 12 is a circuit to test common-mode rejection ratio (CMRR) of a transformer. A wideband leveled RF generator drives both sides of the primary winding. Both sides of the secondary winding are attached to a resistive termination and a wideband RF millivoltmeter. The output is the component of the common-mode signal which "leaks" through the primary-to-secondary capacitance of the transformer. If the transformer has an interwinding shield, it is returned to the ground plane. Figure 13 is a plot of the CMRR vs. frequency obtained with this test, for two different transformer designs. Since improved CMRR is a prime motivation to use a transformer, the CMRR may be the most significant parameter to specify.

The same consideration for suppression of external interference on received signals also apply to reduction of conducted interference emitted by the equipment. The symmetrical nature (reversing input and output) of the interference equivalent circuit means emitted interference such as microprocessor clocks, high speed DSP clocks, etcetra will be reduced by the same ratio of common-mode rejection as demonstrated above. Regulatory compliance for conducted EMI on the digital audio cables and connectors can be improved by using low-capacitance shielded transformers.

7. Common-Mode Interference: Induced Jitter Test

Figure 14 is a test for jitter-induced by high frequency common-mode asynchronous noise. An Audio Precision System 1 generates an AES/EBU test signal output transmitted via one of 4 pairs in a 31m long cable to an AES/EBU decoder circuit.

The cable specification is:

Type	=	Cat 5 Network wire, Belden 1538A
Length	=	31m
Wire	=	#24 ga. PVC twisted pair
Inductance	=	415 uH/conductor
Capacitance	=	1940 pF/each conductor to all others
Resistance	=	2.67 Ω /conductor

Three unused cable pairs are connected in parallel to an RF generator with a 50 Ω termination and a wideband millivoltmeter to monitor the interference level. The interfering signal is applied to three unused pairs, generating crosstalk that appears as common-mode noise current coupled through the capacitance between the unused pairs and the active pair. The receiver circuit couples the digital signal through the transformer under test and decodes it with a Crystal Semiconductor CS8412 AES/EBU receiver.

The rising edges of the frame sync output (pin 11 of CS8412) of the decoder define the output sample time. This clock is analyzed by a Hewlett-Packard 5370B Time Interval counter, capable of statistical analysis and 20 ps jitter measurement. (20 ps is the time for a light ray to cross a pencil's diameter!)

The counter measures each frame sync (Fs) period, collects a sample set of 100 to 10,000 periods and calculates the standard deviation of that set, a direct measure of the wideband RMS jitter. The impact of the jitter on SNR of an audio signal is a function of the frequency and amplitude of the signal, and the architecture of the converter. References [5-8] derive the relationship between jitter and recovered audio's dynamic range.

8. Common-Mode Interference: Induced Jitter Results

Good recovered clock quality is essential for low bit-error-rate data recovery. In addition, low recovered clock-jitter is essential to minimize audio signal degradation for those systems relying on the recovered clock for operating A/D and D/A converters.

Figure 15 is a plot of RMS jitter in nanoseconds vs. common-mode interference level in dBm (ref. 50 Ω) taken with the setup described above. The interfering signal is a 6.98 MHz sinewave. The upper curve is an unshielded transformer with 40 pF of C_{P-S} . The lower curve is a 3 pF C_{P-S} shielded transformer. This data illustrates the dramatic effect that a high CMRR transformer can have on jitter. For optimum performance in high noise environments and low jitter applications, a low capacitance, high CMRR, shielded transformer is required.

9. Transmission System with Shielded Transformers

Common-mode noise on a transformer primary induces a current through the transformer primary-to-secondary capacitance C_{P-S} (typically 10 to 40 pF) and then to the receiver input. Adding an interwinding shield provides a substantial (5 to 20 fold) improvement in CMRR.

Figure 16 shows how a shield introduces a capacitance (from primary to shield, $C_{P-Shield}$) which shunts most of the common-mode current away from the secondary winding and diverts it to ground. Some capacitance remains from primary to secondary (C_{P-S}) due to leads, PCB traces, and other parasitic capacitance. It is possible to realize 1 pF in a good SMD design! The common-mode current divides in proportion to the capacitance ratio, $C_{P-S}/C_{P-Shield}$. The receiver input sees only the small current coupled through C_{P-S} . In general, addition of a shield to any transformer design leads to a tradeoff, increasing leakage inductance and reducing bandwidth.

10. Typical Digital Audio Transformer Applications

10.0 Balanced 110 Ω System with Shielded Transformers

Figure 17 is a balanced 110 Ω system using shielded transformers both on the transmit and receive sides. The shield of each transformer is connected to the ground plane of the associated IC. The connection from shield to the ground plane must have short, low inductance path, to maintain the shield's effectiveness.

10.1 75 Ω Unbalanced Interface with 2:1 Ratio Transformers

Figure 18 is a 75 Ω unbalanced system. Since the transmit IC has 0 to 5 V output, the transformer has a 2:1 step-down ratio to match the 300 Ω primary impedance to the 75 Ω cable impedance and reduce the output voltage level to 625 mV. The series 0.1 μ F capacitor blocks the dc bias at the output of the IC to prevent transformer core saturation. The transformer shield returns to the transmit IC ground pin and ground plane. The secondary goes to the connector, with the low side of the secondary connected to the connector shell.

The shell of an unbalanced connector may either float or go to chassis (earth) ground. If a floating connector shell is used (e.g. to break ground loops) then a small high-frequency RF return capacitor for example 10 to 100 pF, may be added from connector shell to chassis (earth) ground.

Any digital audio receiver IC has finite dynamic range and CMRR. Low level, unbalanced signals, (e.g. SPDIF from a weak source on a lossy cable) will increase the error rate and jitter since the receiver has less margin and interference rejection. A step-up transformer can increase the signal voltage level, improving both receiver performance and common-mode rejection.

On the receive side of Figure 18, the 75 Ω connector feeds a transformer with a 1:2 step-up. The secondary is terminated with 300 Ω . The transformer shield is attached through a short and direct path to the ground pin of the receive IC (e.g. at local bypass capacitor, and ground plane). The same considerations mentioned above for transmit side floating connectors also apply to the receiver.

10.2 Bridging Unbalanced Input and Floating Coax Connector

Figure 19 shows a high impedance bridging receiver using a 1:2 step-up transformer. The 3.9 k Ω resistor terminates the transformer to maintain proper frequency response. That resistor is reflected to the primary as 975 Ω , presenting a light load to the 75 Ω line. The connector is shown floating, with the RF bypass capacitor to chassis ground.

10.3 Phantom Power Remote Digital Device

Low-power A/D and D/A converters make remotely powered devices such as digital microphones a reality. Figure 20 is a phantom power circuit where the signal cable carries the DC power. The power source on the local side of the system is connected to the center tap of the output transformer. The remote device gets DC power from the input transformer's center tap. Decoupling filters should be used on both ends.

10.4 Dual output transmitter, 110 Ω Balanced and 75 Ω Unbalanced

Some applications require both balanced and unbalanced outputs. Figure 21 uses a 1:1 center tapped transformer. The 110 Ω primary resistor reflects to the center tap of the secondary as 27 Ω . The 47 Ω series resistor matches the 75 Ω unbalanced output impedance. The balanced 110 Ω output is obtained across the entire secondary. Note that only one of the two outputs may be connected at one time.

11. Comparison of Typical Transformers

Figure 22 compares the parameters and performance of seven types of commercially-available digital audio transformers. Note the substantial differences in most parameters. The higher cost of the best available transformer is easily justifiable in high-quality equipment designs, since that cost is a negligible fraction of the total material cost, and the benefits are substantial. The worst transformers can degrade performance and specifications of the equipment and impair regulatory compliance.

12. PC Layout Considerations

The transformer isolates the relatively noisy external connections from the (hopefully!) clean internal circuitry of the equipment. The low capacitance and shield of the transformer provide rejection of high-frequency common-mode interference present on the external connection. Good RF printed circuit layout practice can further improve the noise rejection of the finished design, for example, by adding ground planes for shielding and by minimizing the primary to secondary capacitance.

A typical PCB layout for a transformer circuit (either input or output) is shown in Figure 23. The external coaxial connector is connected with short and direct traces to the transformer. The traces to the IC may be longer, but they should be kept away from the connectors. Two ground planes are used, which are split under the transformer. The ground plane facing the connector goes to chassis (earth) ground. The ground plane facing the IC and the transformer shield (if present) are attached to the ground return pin of the associated IC with a short and direct trace.

CONCLUSIONS

Transmission circuits for digital audio signals are improved by the use of transformers on both the transmit and receive sides. Measurements show that the transformer has substantial impact on the digital audio signal waveform, the rejection of common mode noise interference and the recovered clock jitter. Good transformer implementation and application can improve signal waveform fidelity, increase EMI rejection and reduce conducted EMI emission. The benefits more than compensate for the small cost increment of high-performance transformers.

ACKNOWLEDGMENTS

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The Effect of Transformers on Transmission of Digital Audio Signals

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Highlights

- Why use transformers in digital transmission?
- Spectrum of digital audio signal
- Transformer effects on waveform
- Transformer effects on noise rejection
- Common-mode rejection: relationship to jitter
- Jitter Histograms for various transformers
- Transformer shielding and noise rejection
- Applications: balanced system with shielded transformers, unbalanced interface,
- high-impedance bridging input, phantom power, dual balanced and unbalanced output
- Comparison of available transformers

Fig. 1 Direct Coupled Transmission System

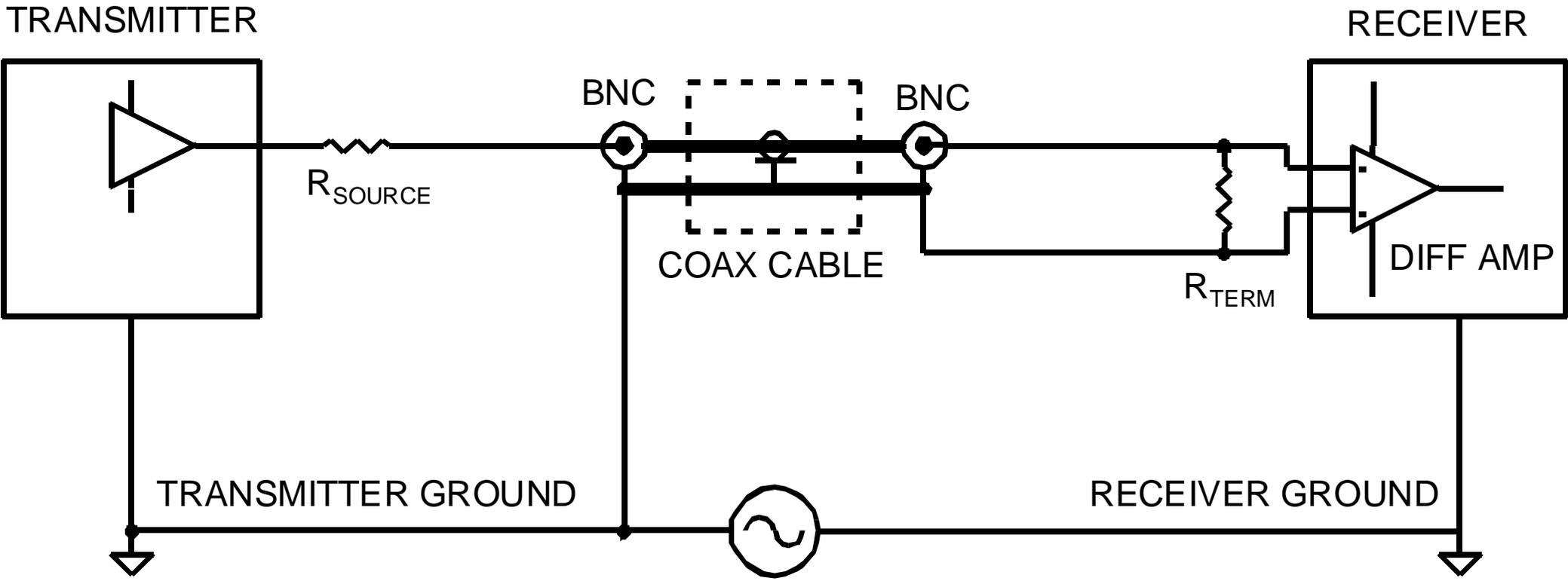
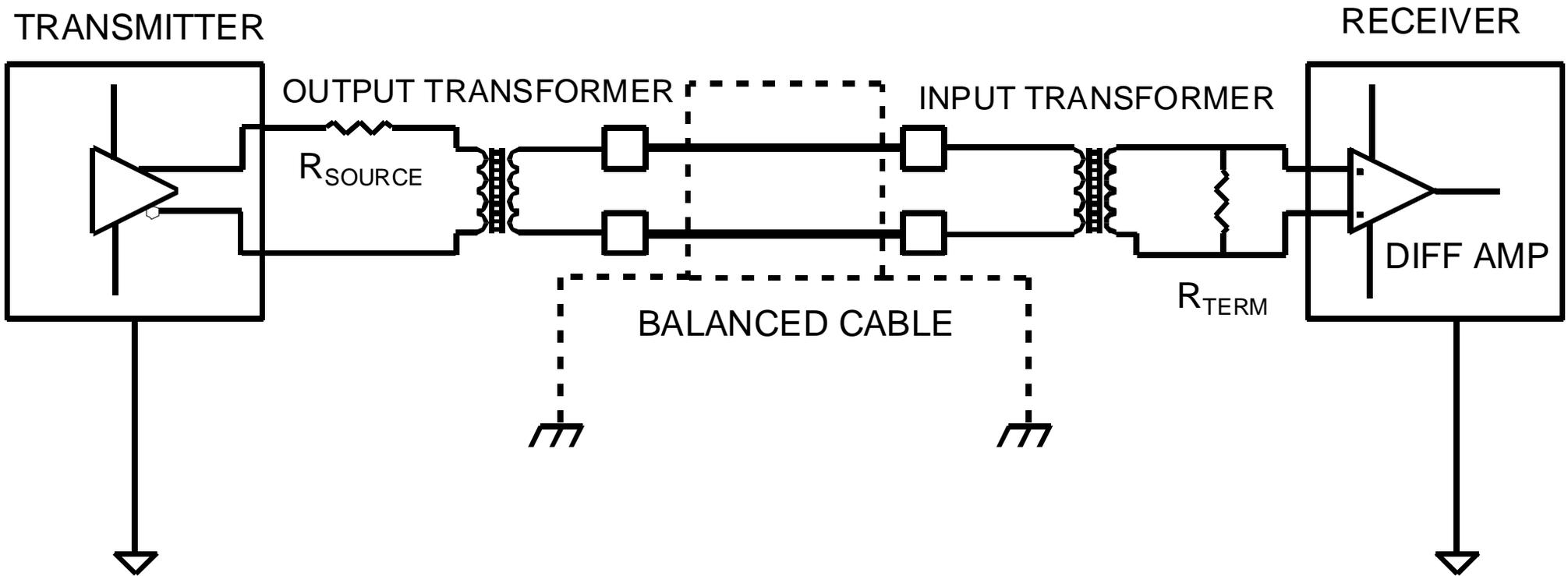


Fig. 2 Transformer Coupled Digital Audio Transmission System



Reasons for Using Transformers in Digital Audio Transmission

- Greatly improve common-mode noise rejection
- Reduce conducted EMI emission to ensure regulatory compliance
- Reduce transmit and receive circuit susceptibility to EMI interference
- Break ground loops
- Balance input and output
- Match impedance and levels
- Achieve lowest possible recovered clock jitter

Fig. 3 Spectrum of Digital Audio Signal
SPDIF 48kHz, 10kHz/div., 10 dB/div.

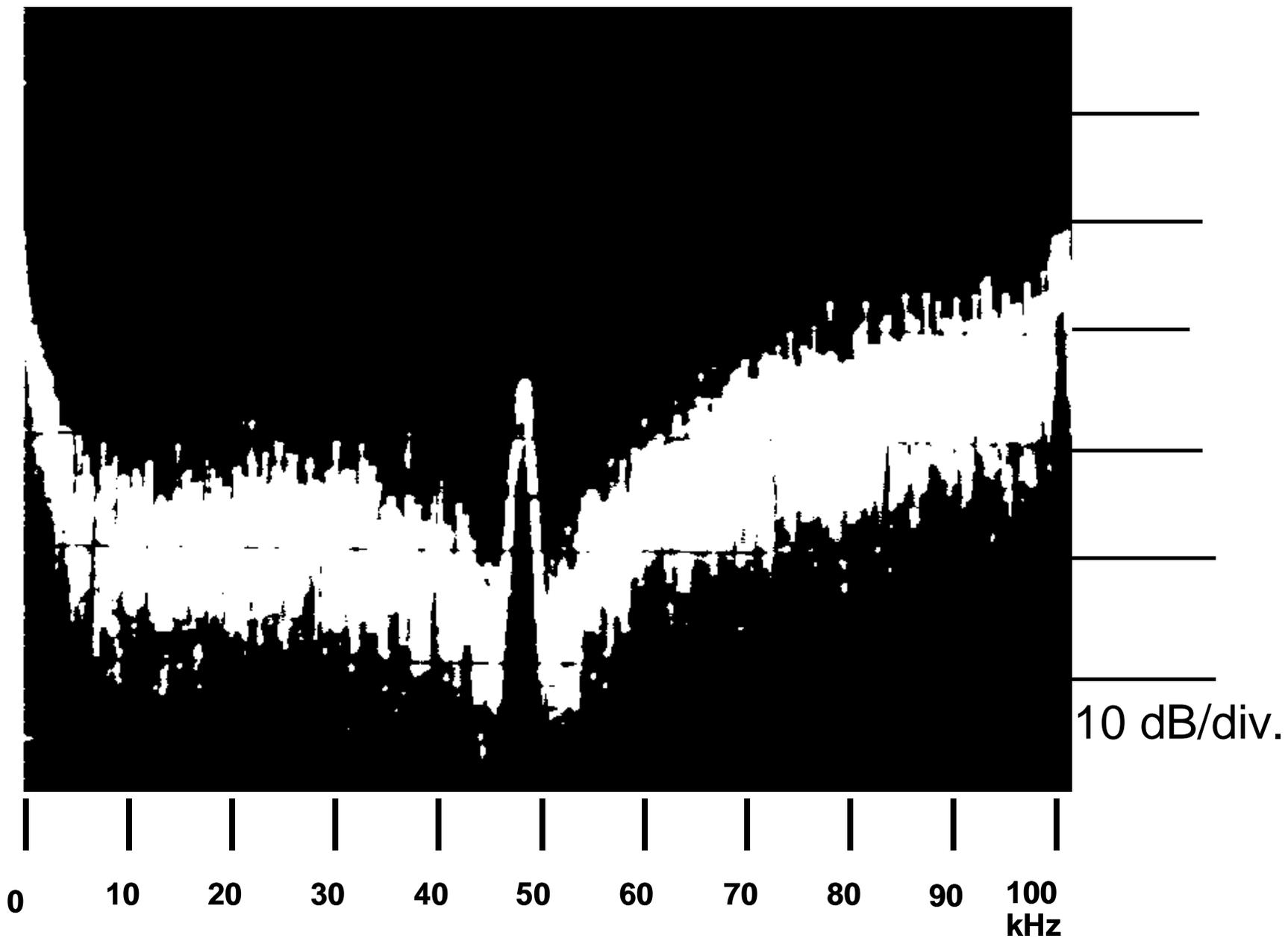


Fig. 4 Spectrum of Digital Audio Signal
SPDIF 48kHz, 5MHz/div., 10 dB/div.

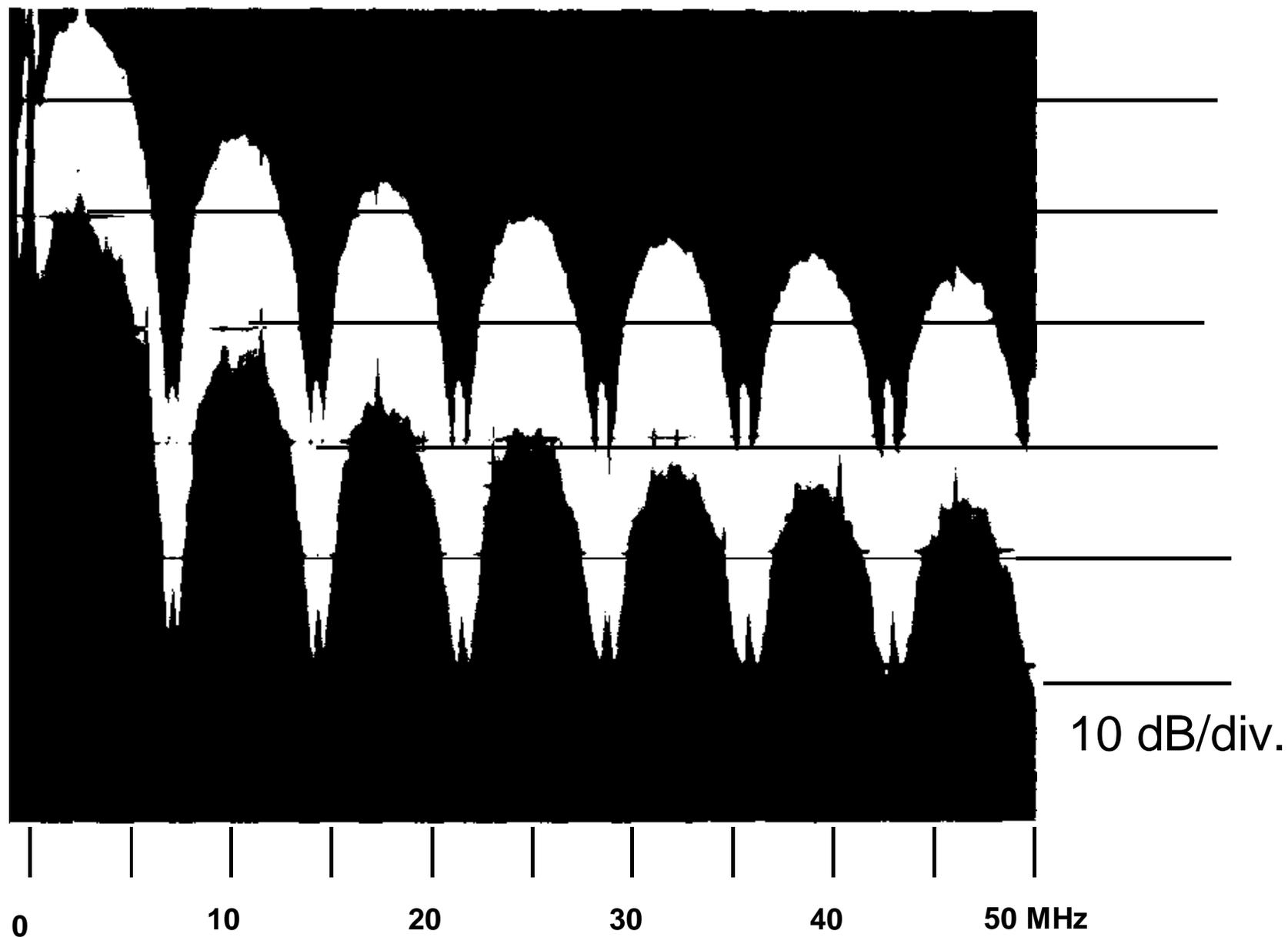


Fig. 5 Equivalent Circuit of Transformer

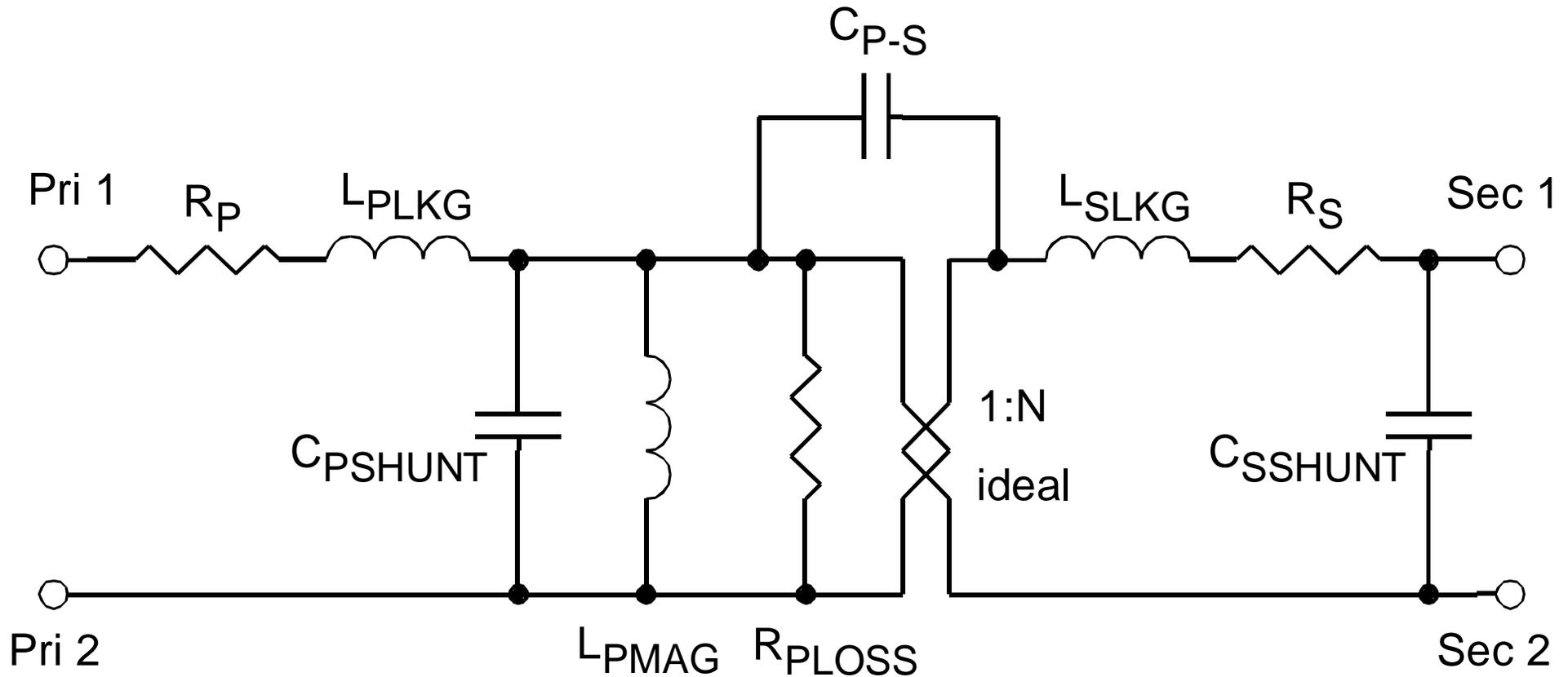


Fig. 6 Frequency Response of Transformer

Parameter: primary inductance, L_{PMAG}

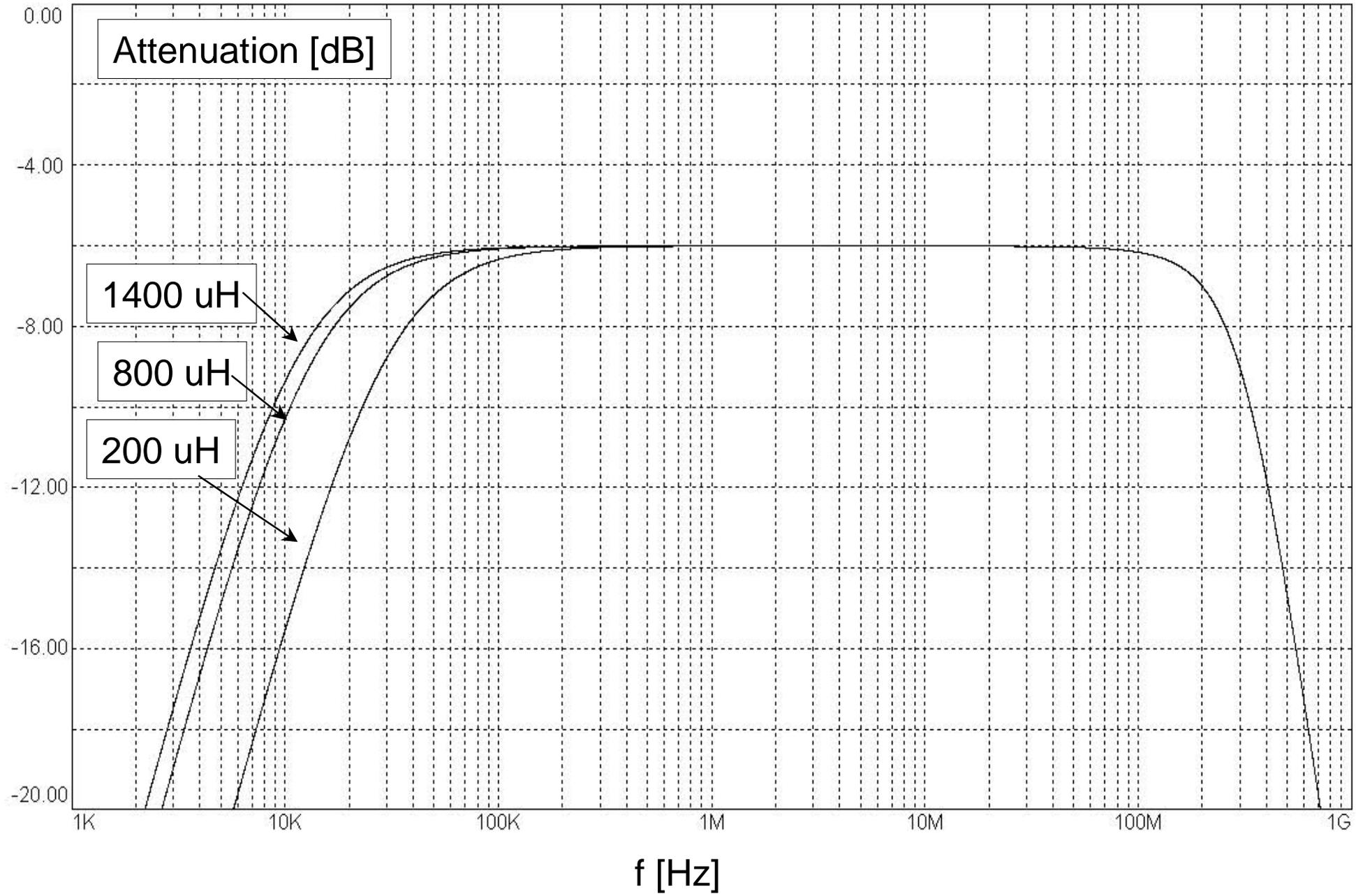


Fig. 7 Frequency Response of Transformer

Parameter: leakage inductance, L_{PLKG}

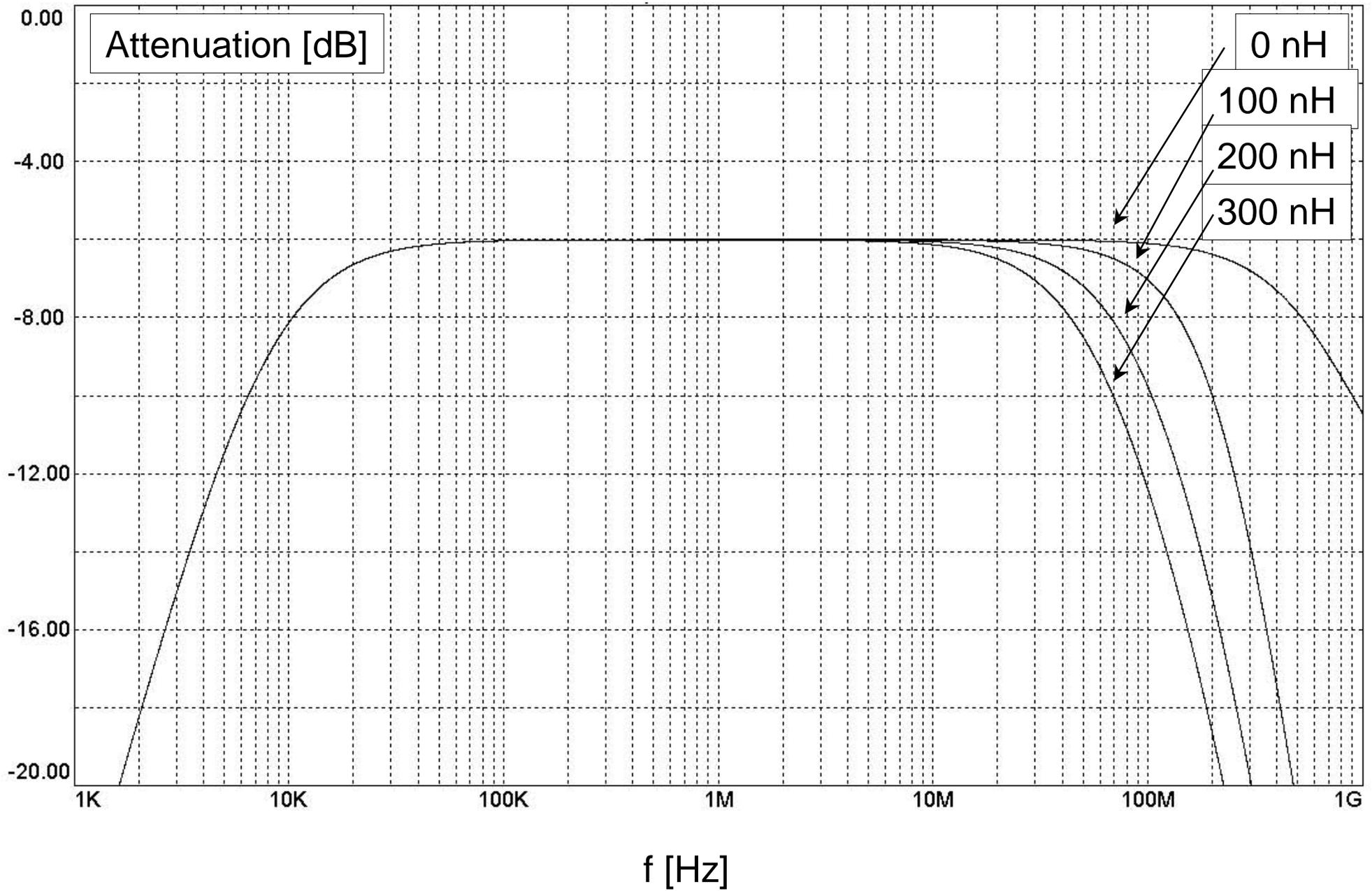


Fig. 8 Effect of Transformer High-Frequency Bandwidth on Pulse Response

12.288 MHz
Pulse Generator

$F_{\text{HIGH}} = 100 \text{ MHz}$
Transformer "S"

$F_{\text{HIGH}} = 35 \text{ MHz}$
Transformer "X"

1 V/div

10 ns/div.

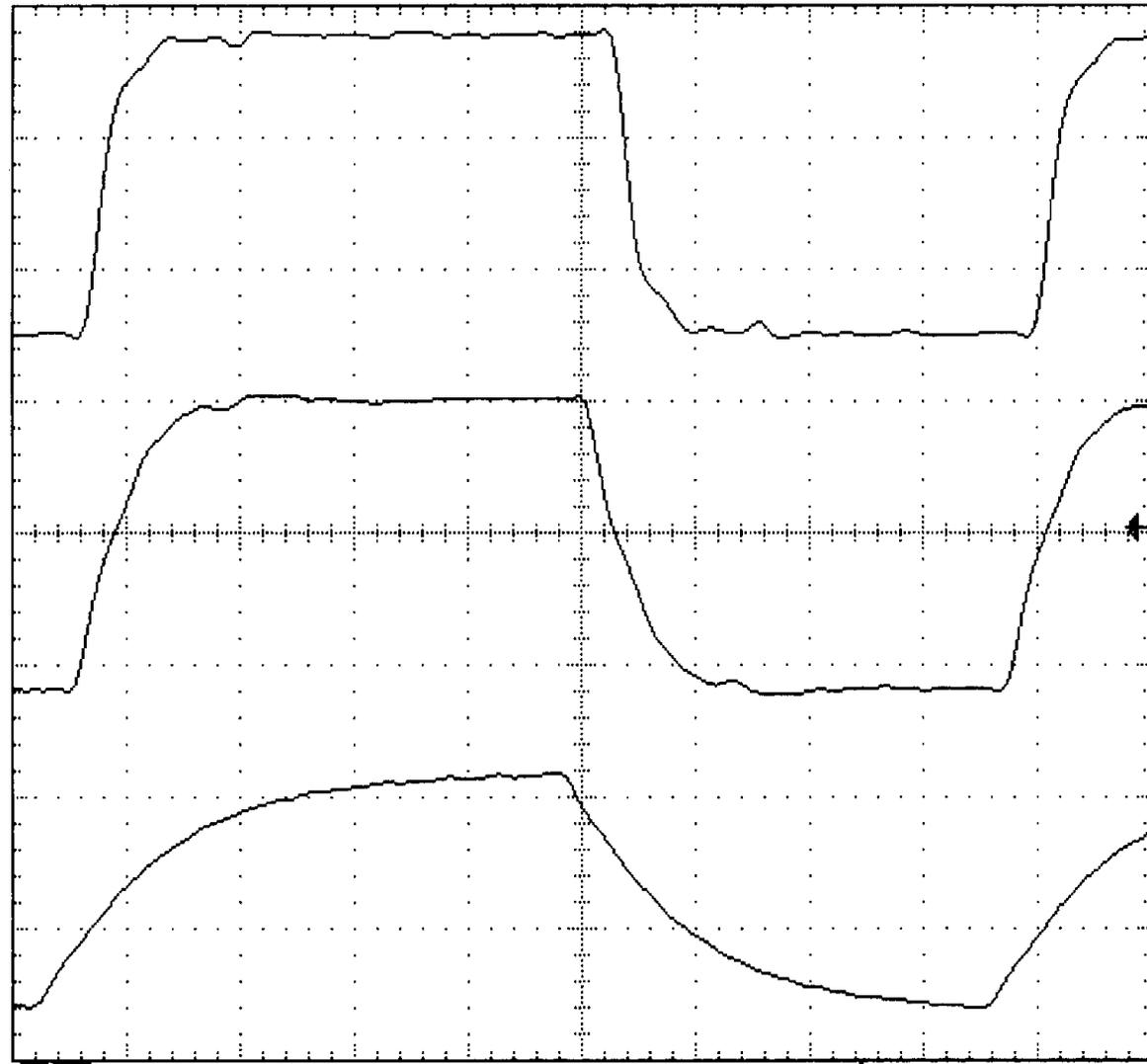
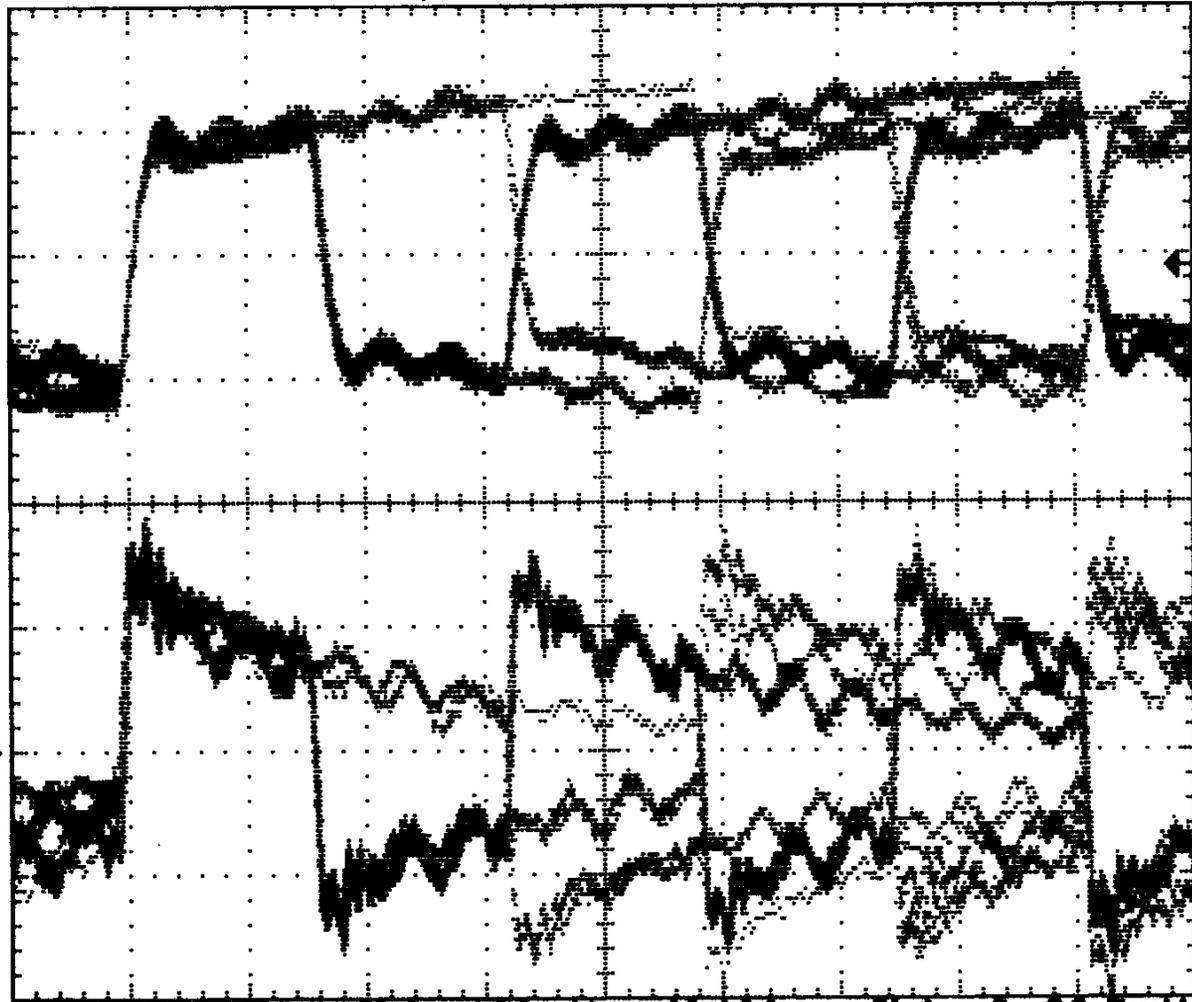


Fig. 9 Effect of Transformer Low-Frequency Bandwidth on Eye Pattern

$F_{\text{LOW}} = 4 \text{ kHz}$
Transformer "S"

$F_{\text{LOW}} = 400 \text{ kHz}$
Transformer "X"

1 V/div



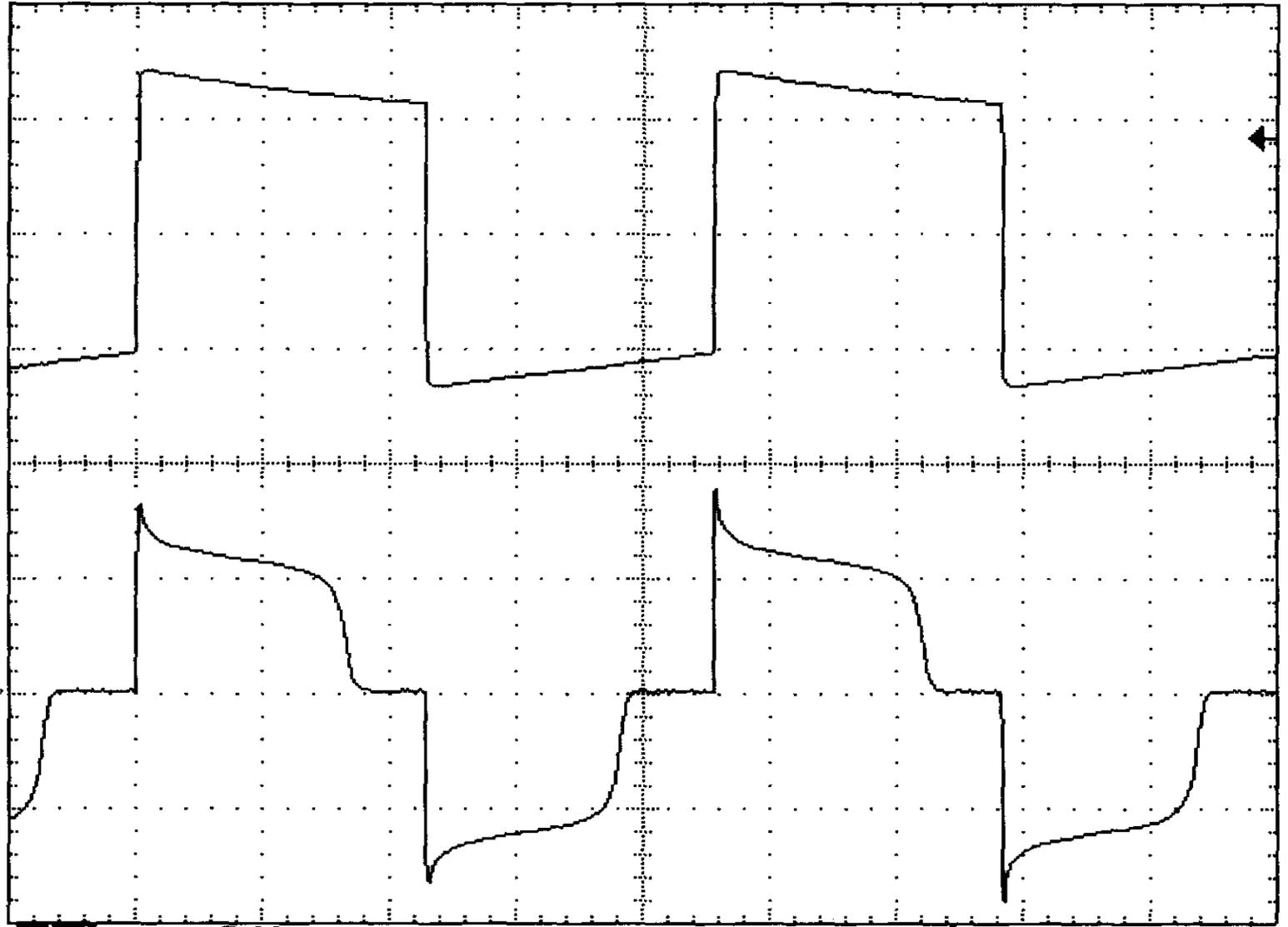
100 ns/div

Fig. 10 Effect of Core Saturation on Waveform

300 μ Vs
Transf. "S"

50 μ Vs,
Transf. "X"

1 V/div



25 μ s/div

Fig. 11 Comparison of Pulse Aberration

Squarewave Generator 12.288 MHz, 75 Ω in/out

Type "S"
very low
aberration

Type "X"
excessive
aberration

1 V/div



5 ns/div.

Fig. 11A Comparison of Aberration

Received AES/EBU 31M UTP 96kHz FS, 110 Ω in/out

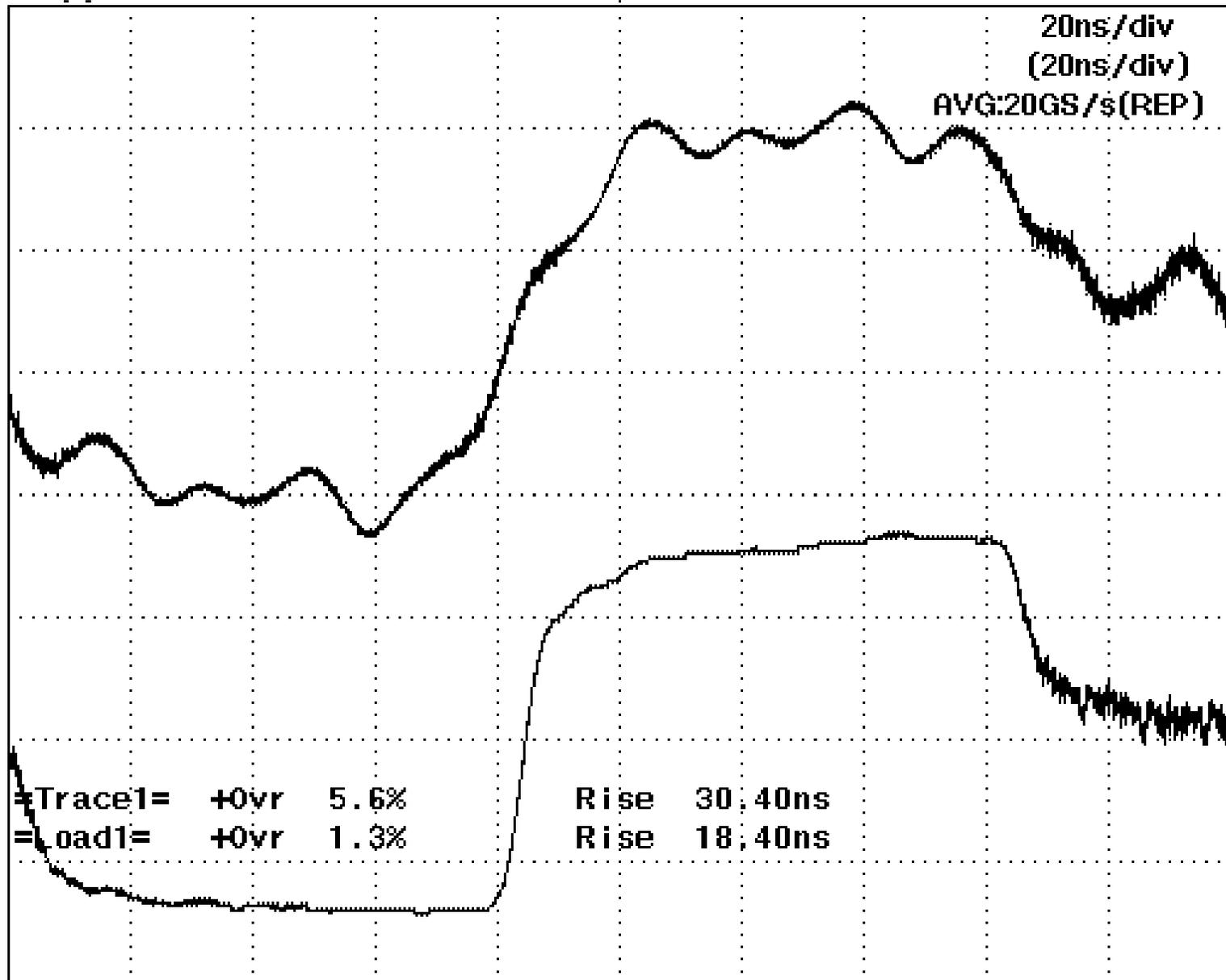


Fig. 12 Common-Mode Noise Rejection Test Circuit

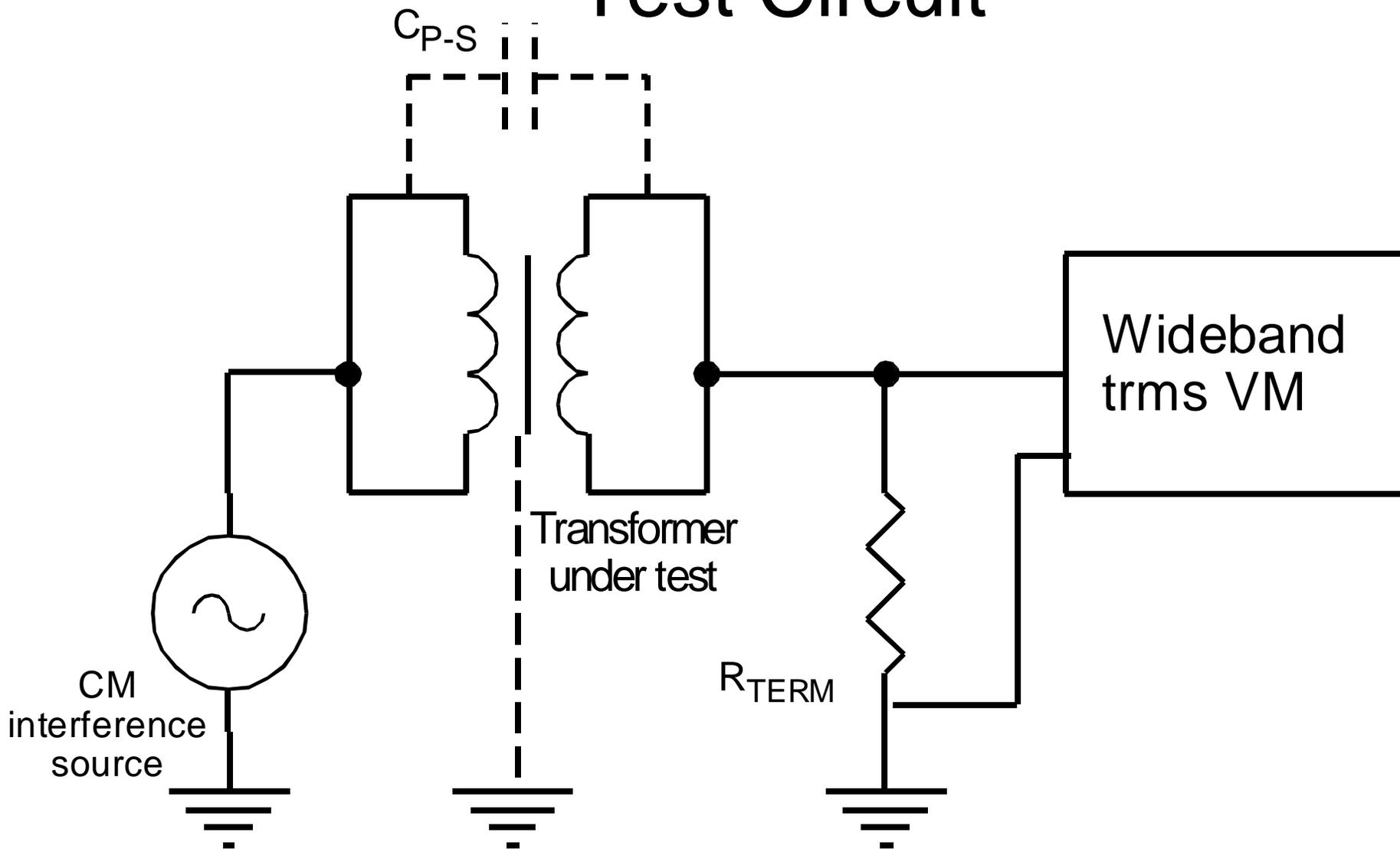


Fig. 13 Common-Mode Noise Rejection

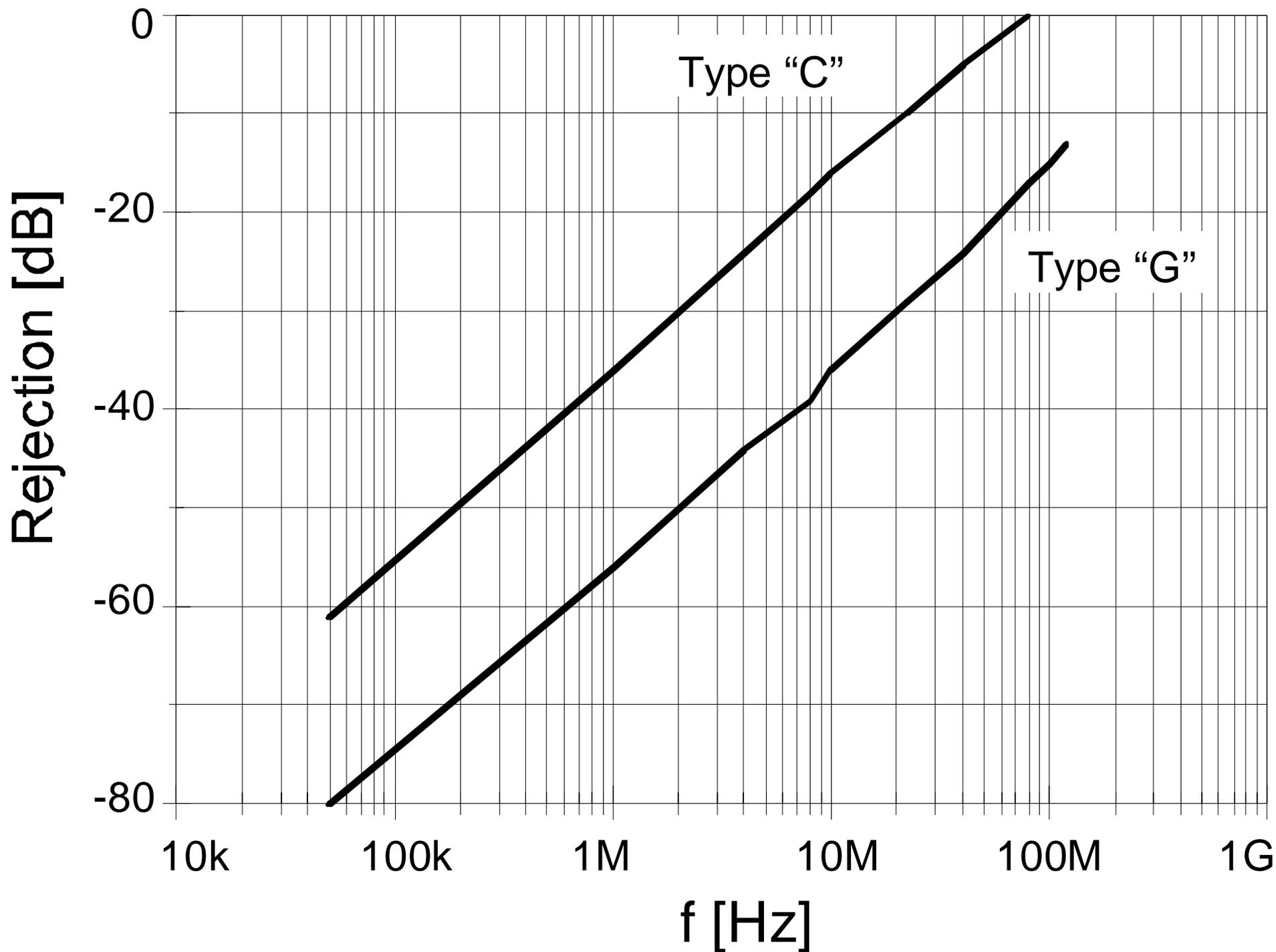


Fig. 14 AP-1/CS8412/48kHz Common-Mode Noise Induced Jitter Test Setup

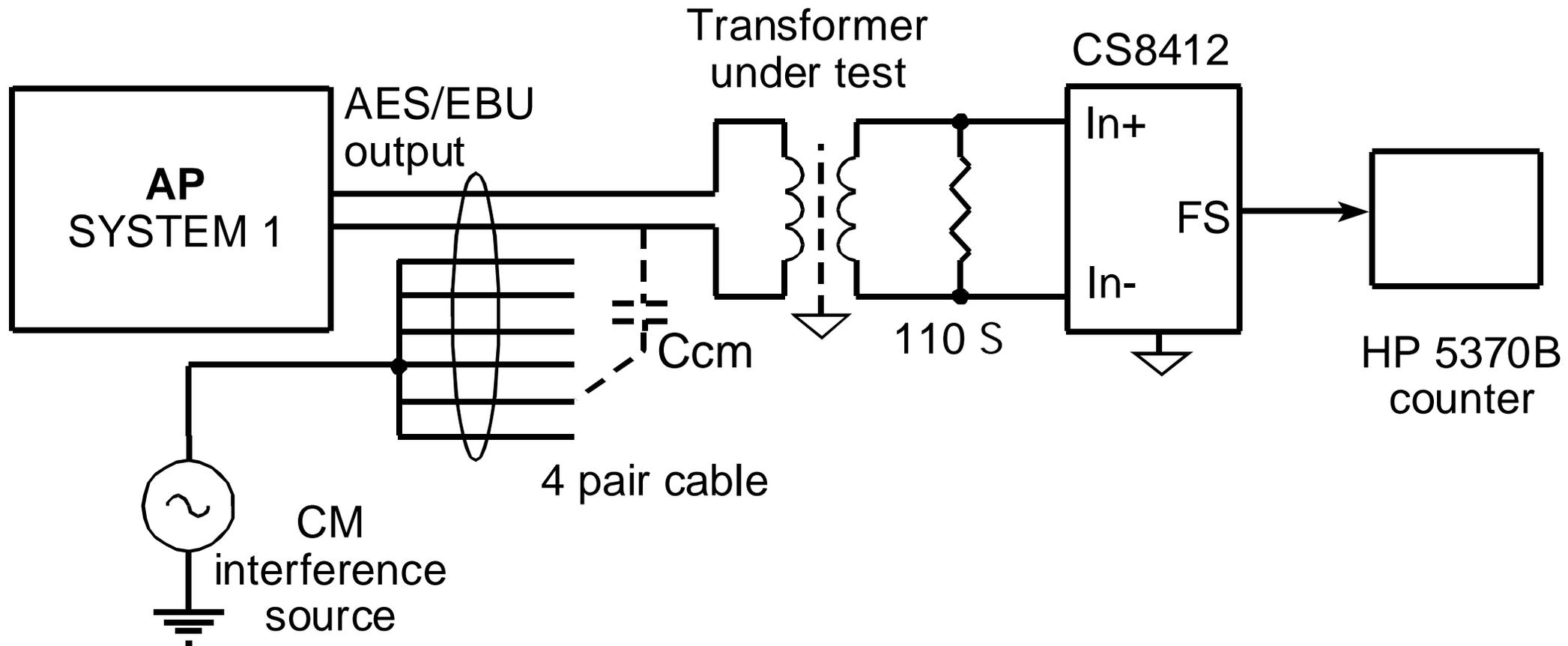


Fig 14 HP 5370B Time Interval Counter

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ELECTRONIC COUNTERS

Universal Time Interval Counter

Model 5370A

- 20 ps single shot time interval counter
- Statistics
- Automatic calibration of systematic errors
- Positive or negative time intervals
- Frequency and period to 100 MHz

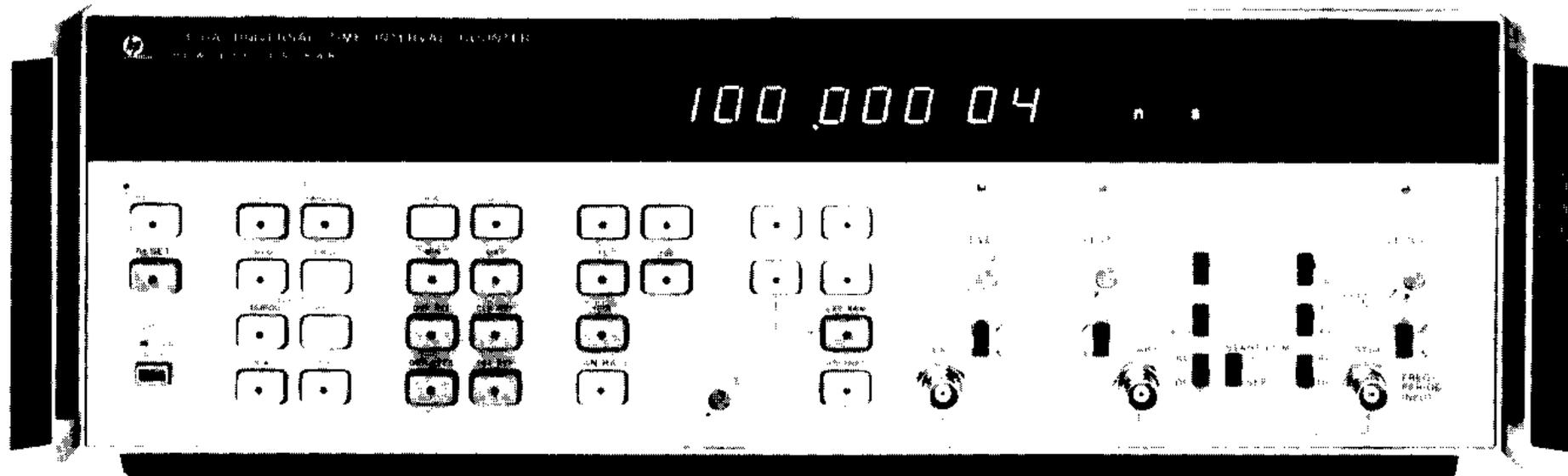


Fig. 15 Induced Common-Mode Noise Jitter Comparison

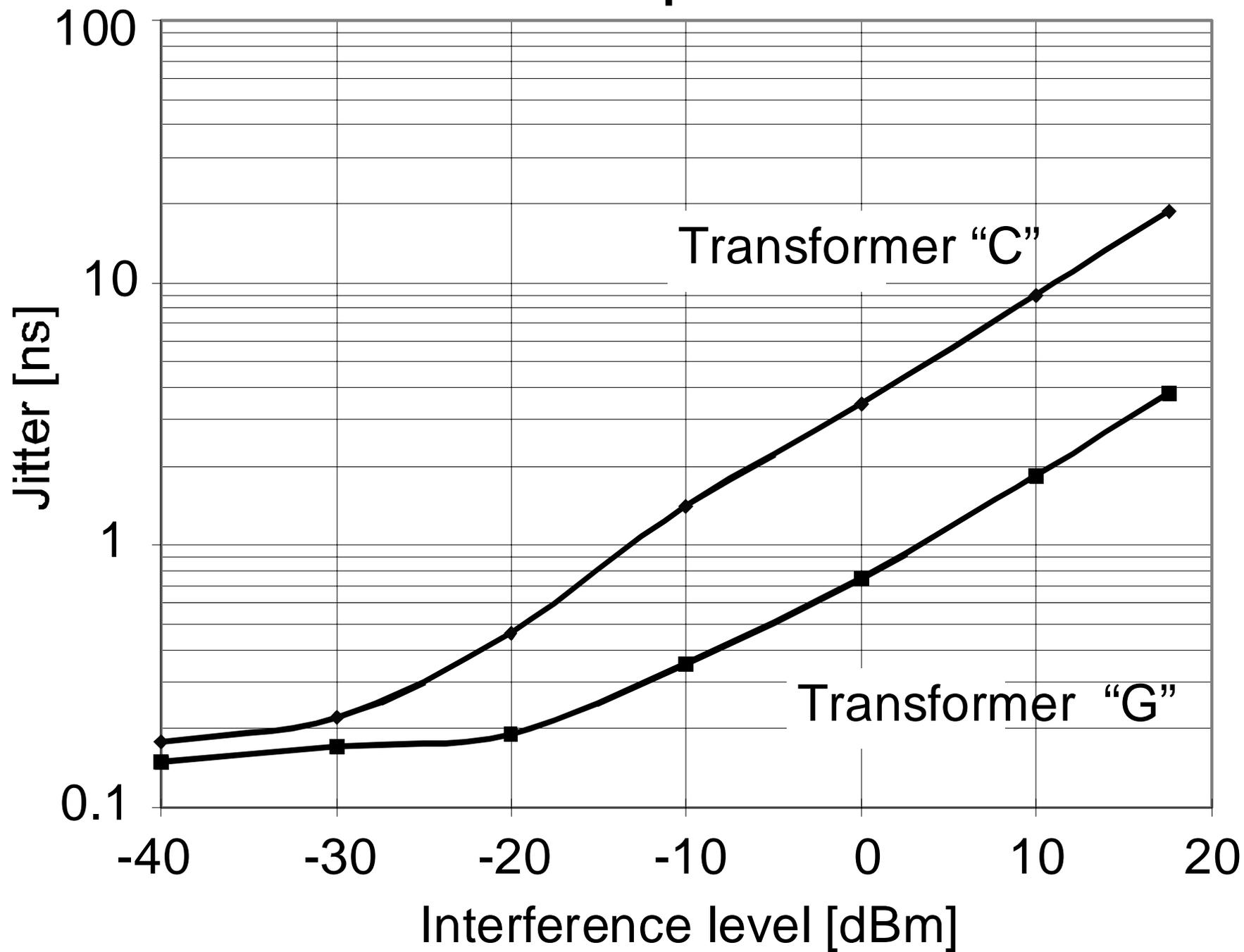


Fig. 14A 96kHz Common-Mode Noise CS8404A/CS8414/96kHz Jitter

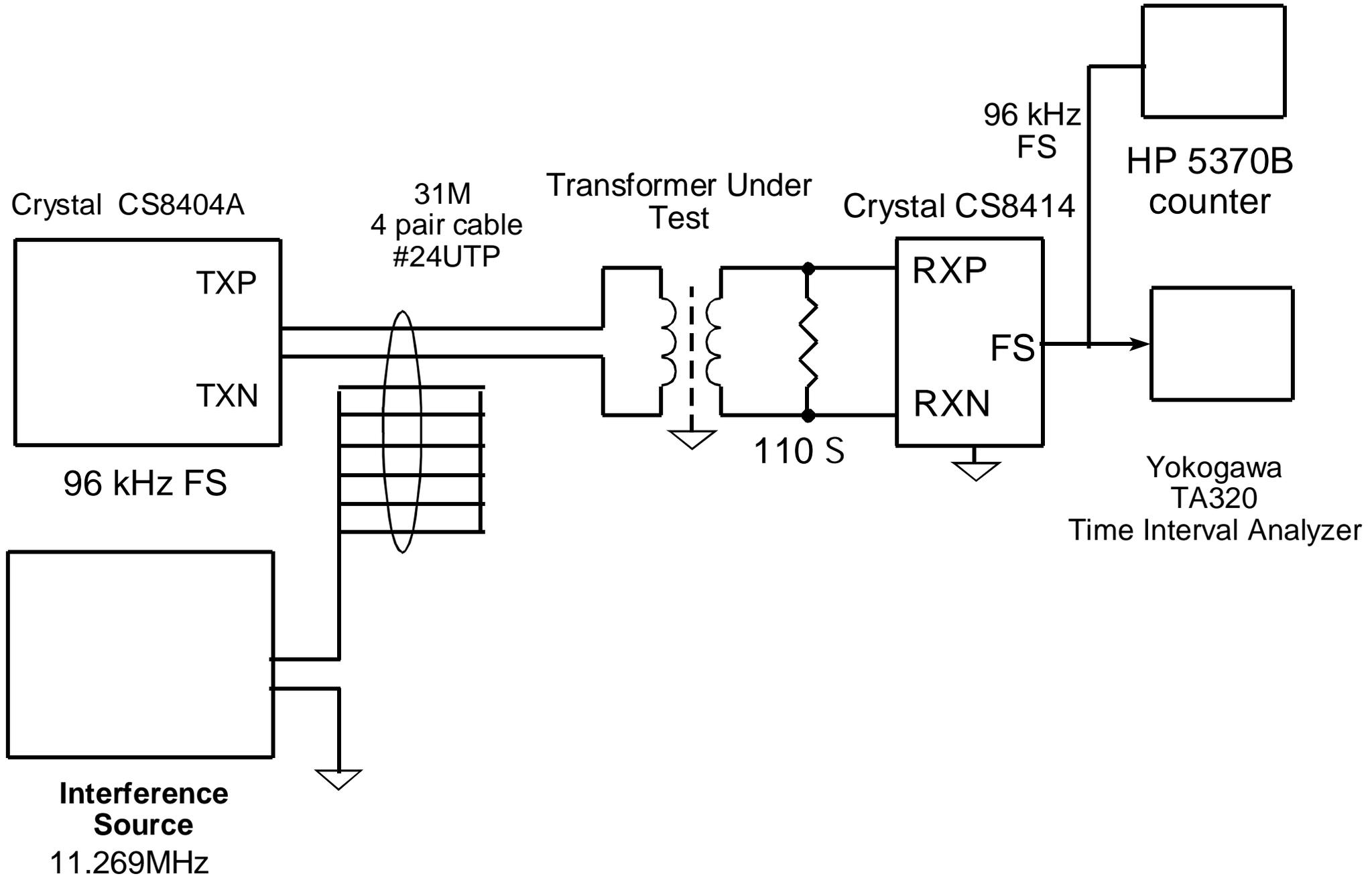


Fig 14A Yokogawa TA320

Time Interval Analyzer

14 MS/s 100 ps resolution

statistics: P-P, std dev, jitter histograms

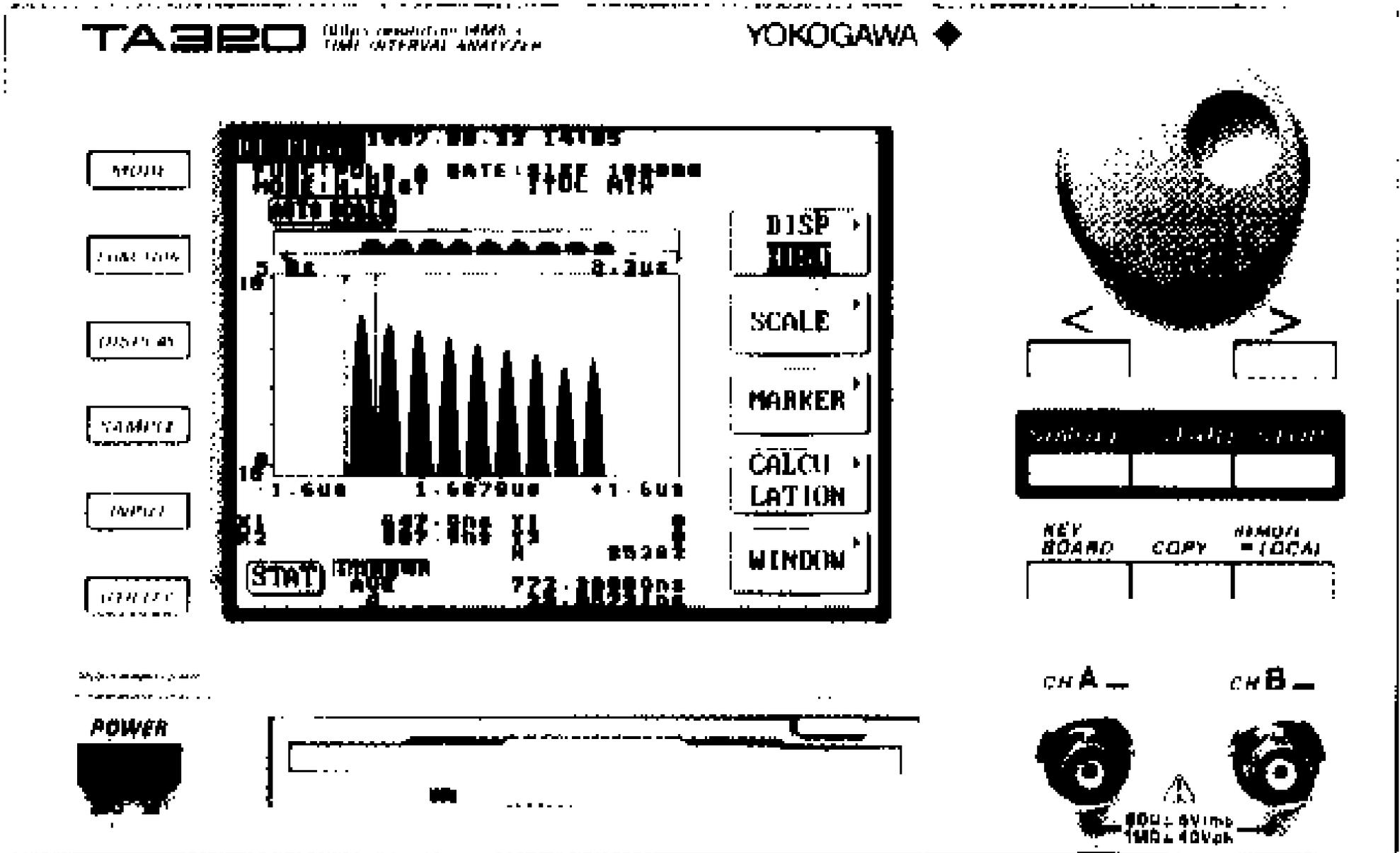


Fig. 14B no trsf.
RMS jitter 3401ps

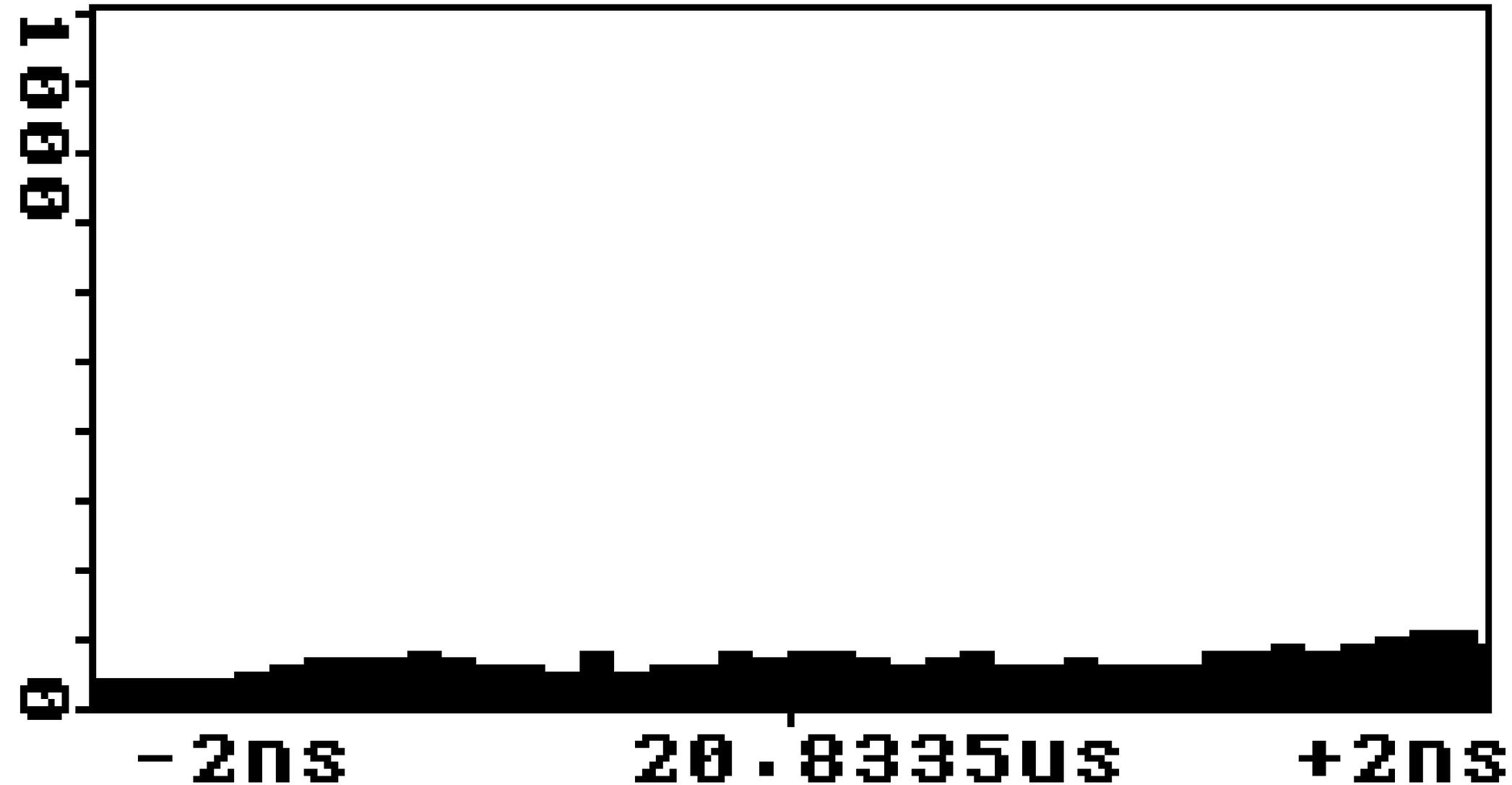


Fig. 14C type "B"
RMS jitter 1540ps

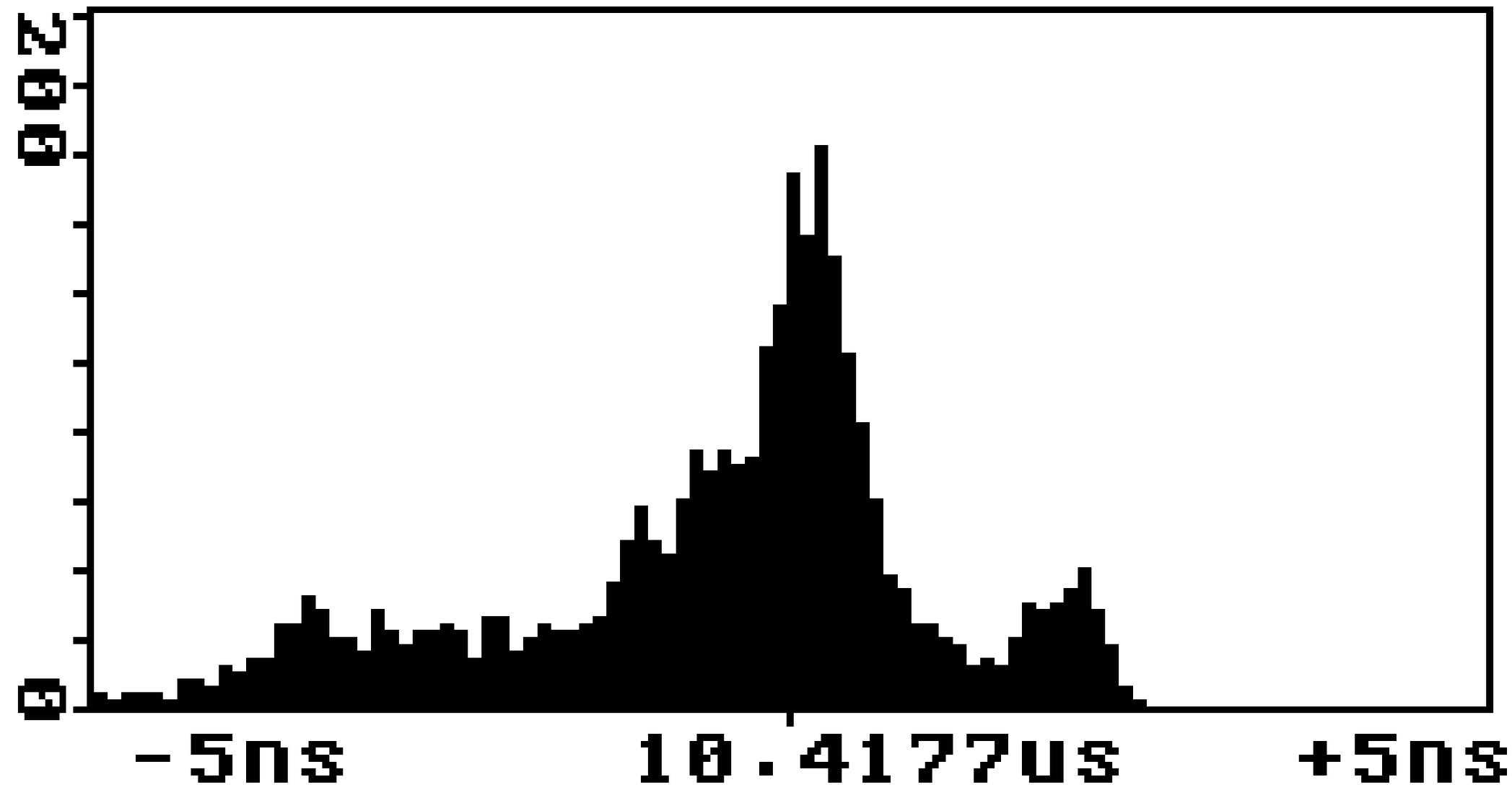


Fig. 14D type "V"
RMS jitter 906ps

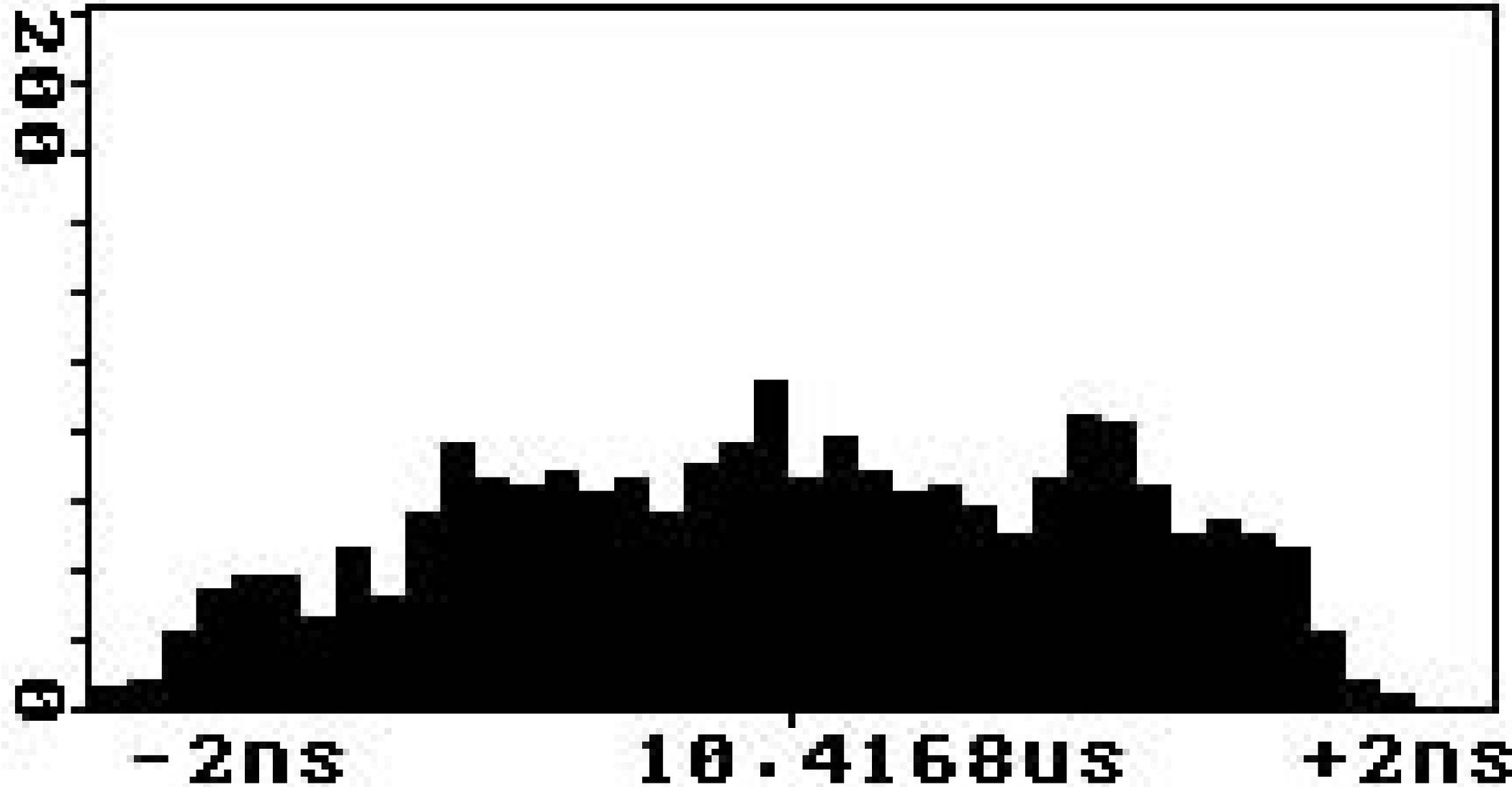


Fig. 14E type "H"
RMS jitter 396ps

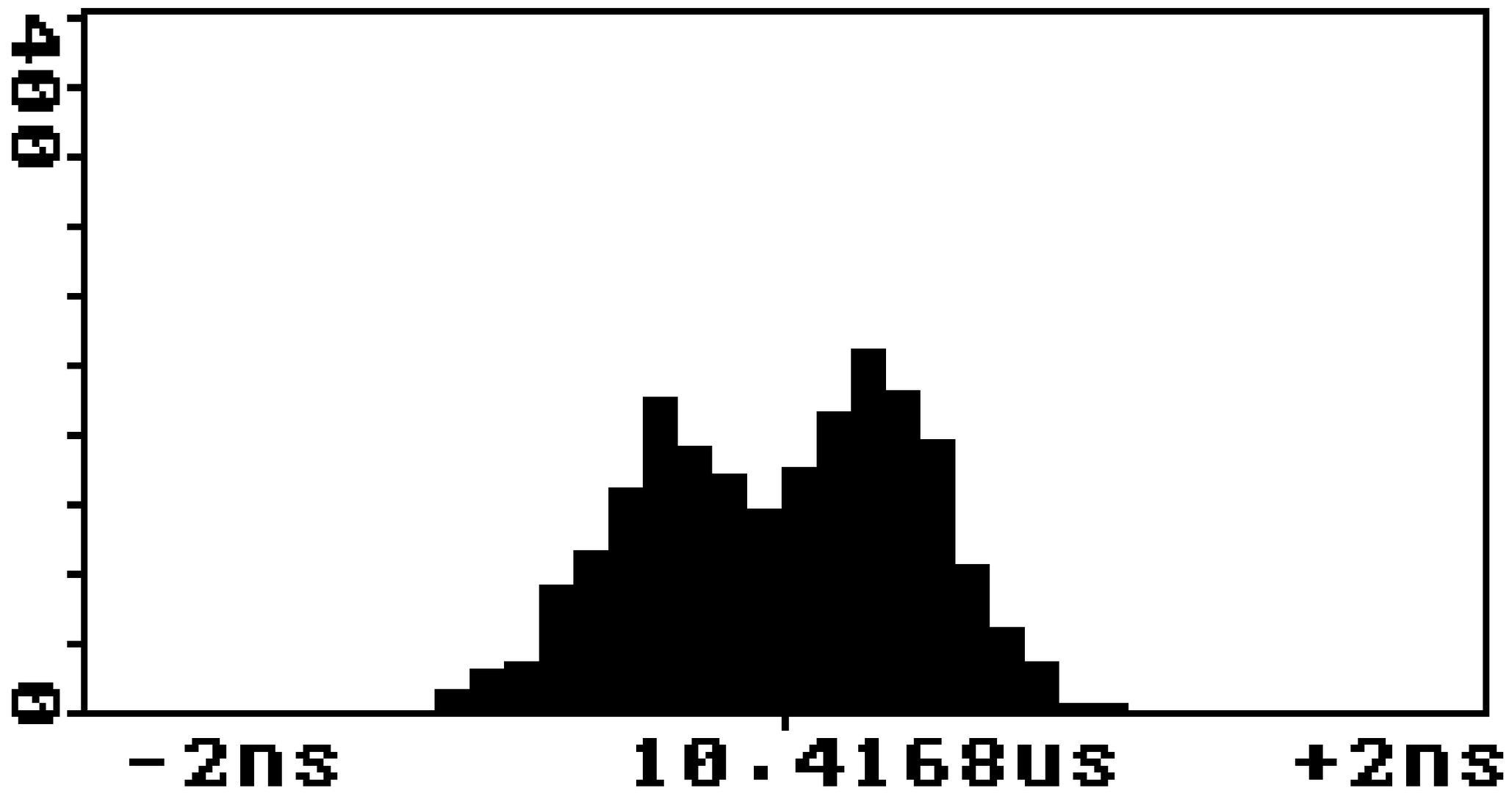


Fig. 16 Shielded Transformer in Common-Mode Noise Equivalent Circuit

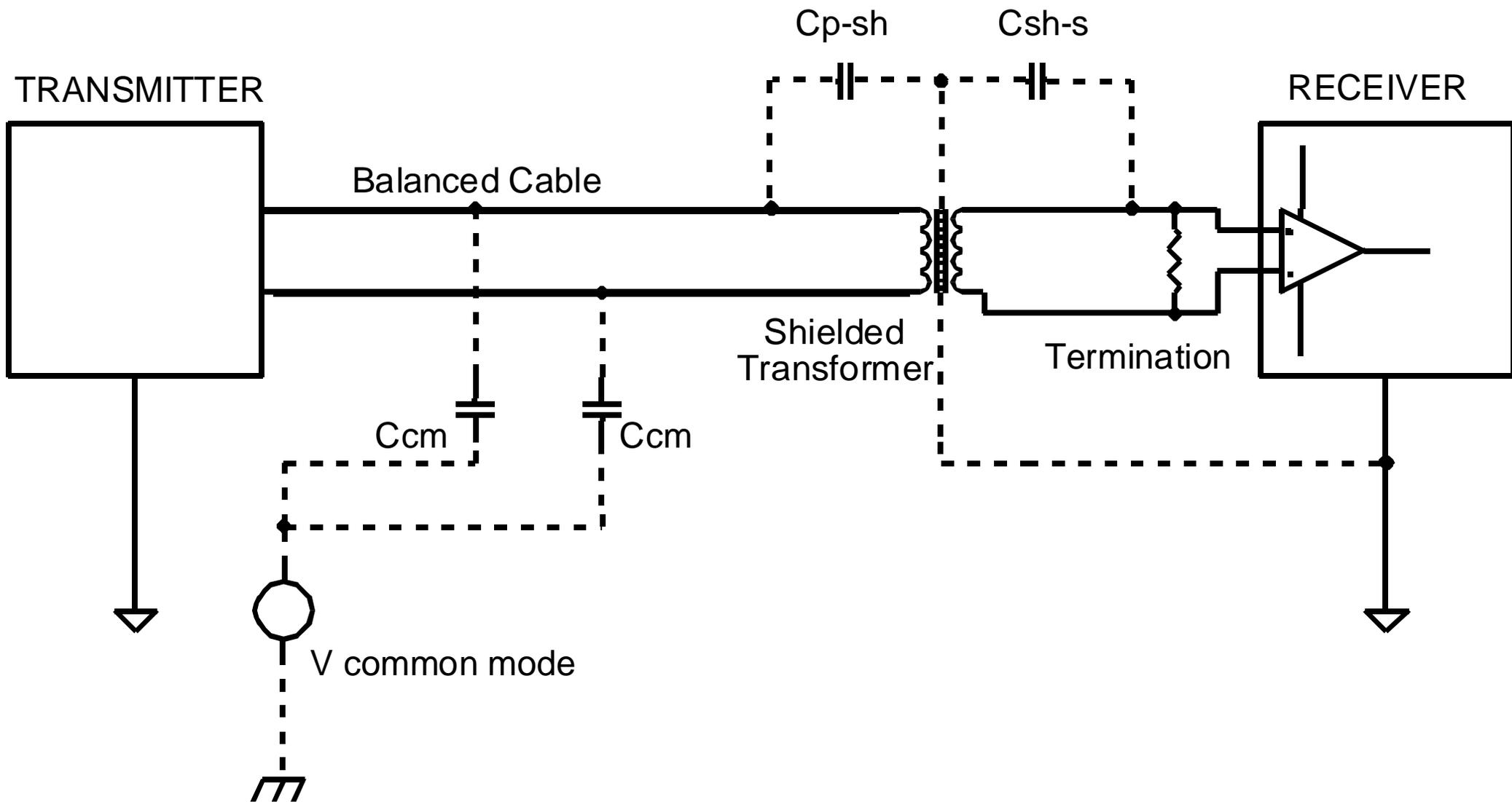


Fig. 17 Balanced Transmission System Using Shielded Transformers

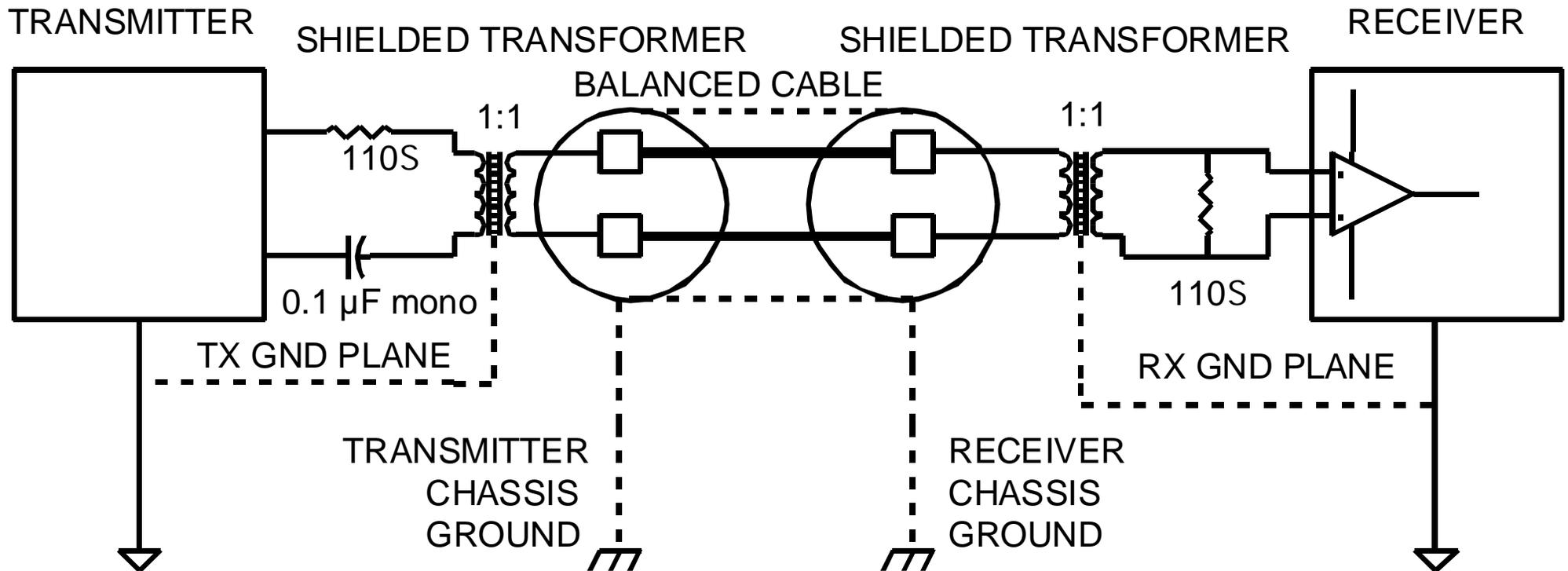


Fig. 18 Application of 2:1 Transformers to 75 Ω Unbalanced System

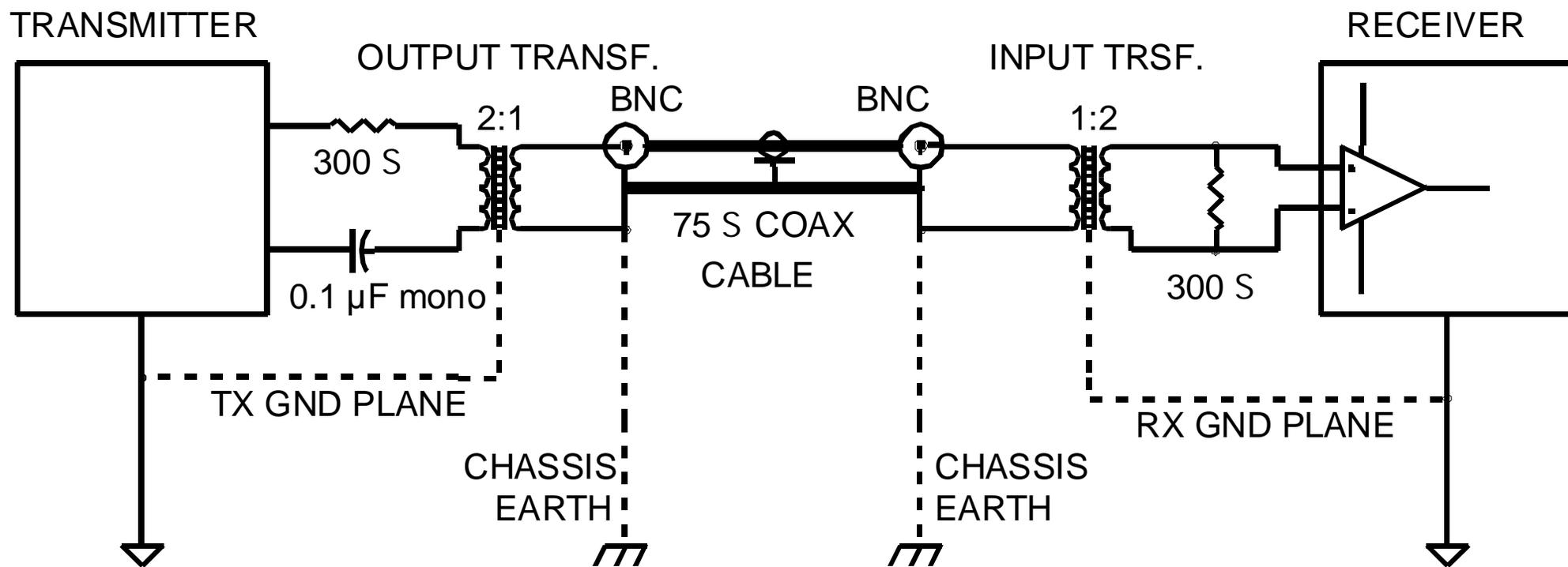


Fig. 19 Hi-Z Bridging Unbalanced Input with Floating Coax Connector

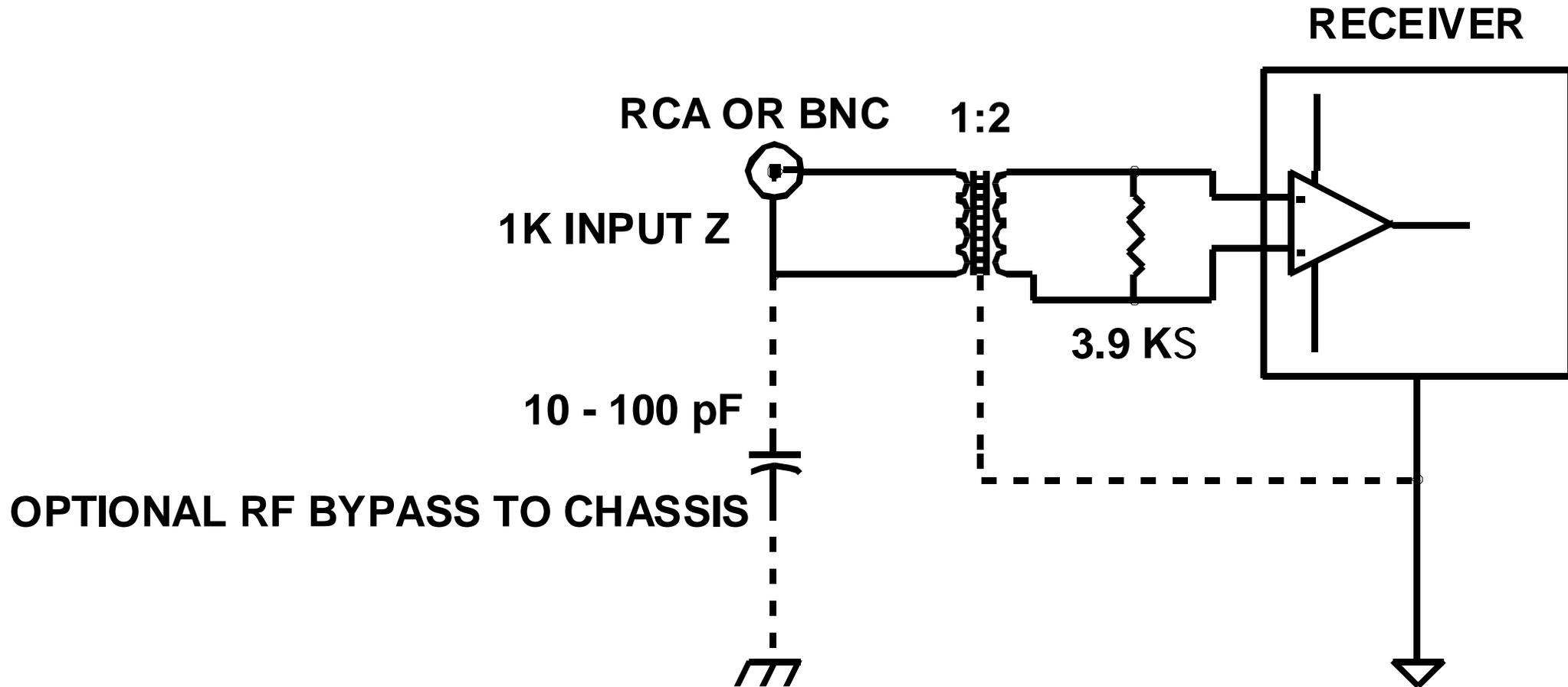


Fig. 20 Phantom Power to Remote Digital Device

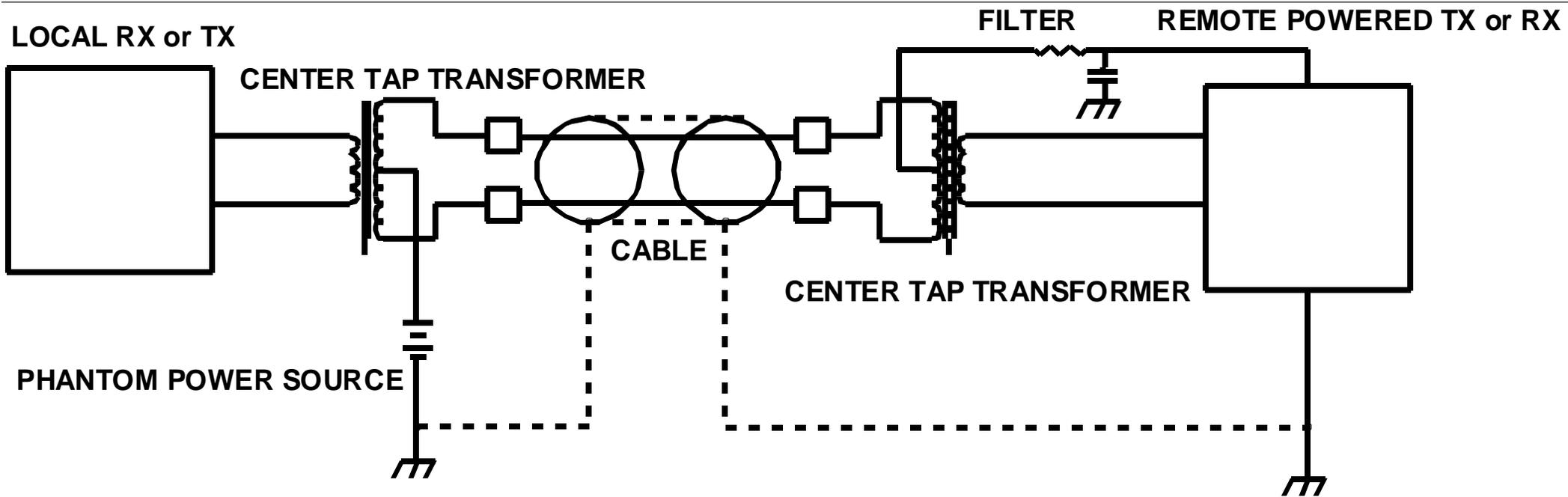


Fig. 21 Dual Output Transmitter
110 Ω Balanced / 75 Ω Unbalanced

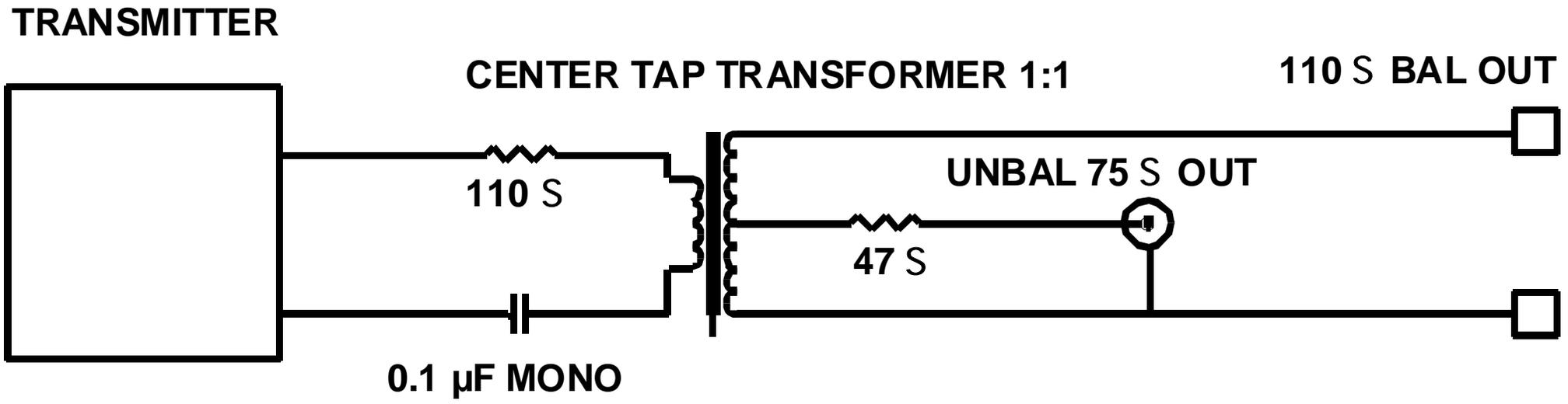
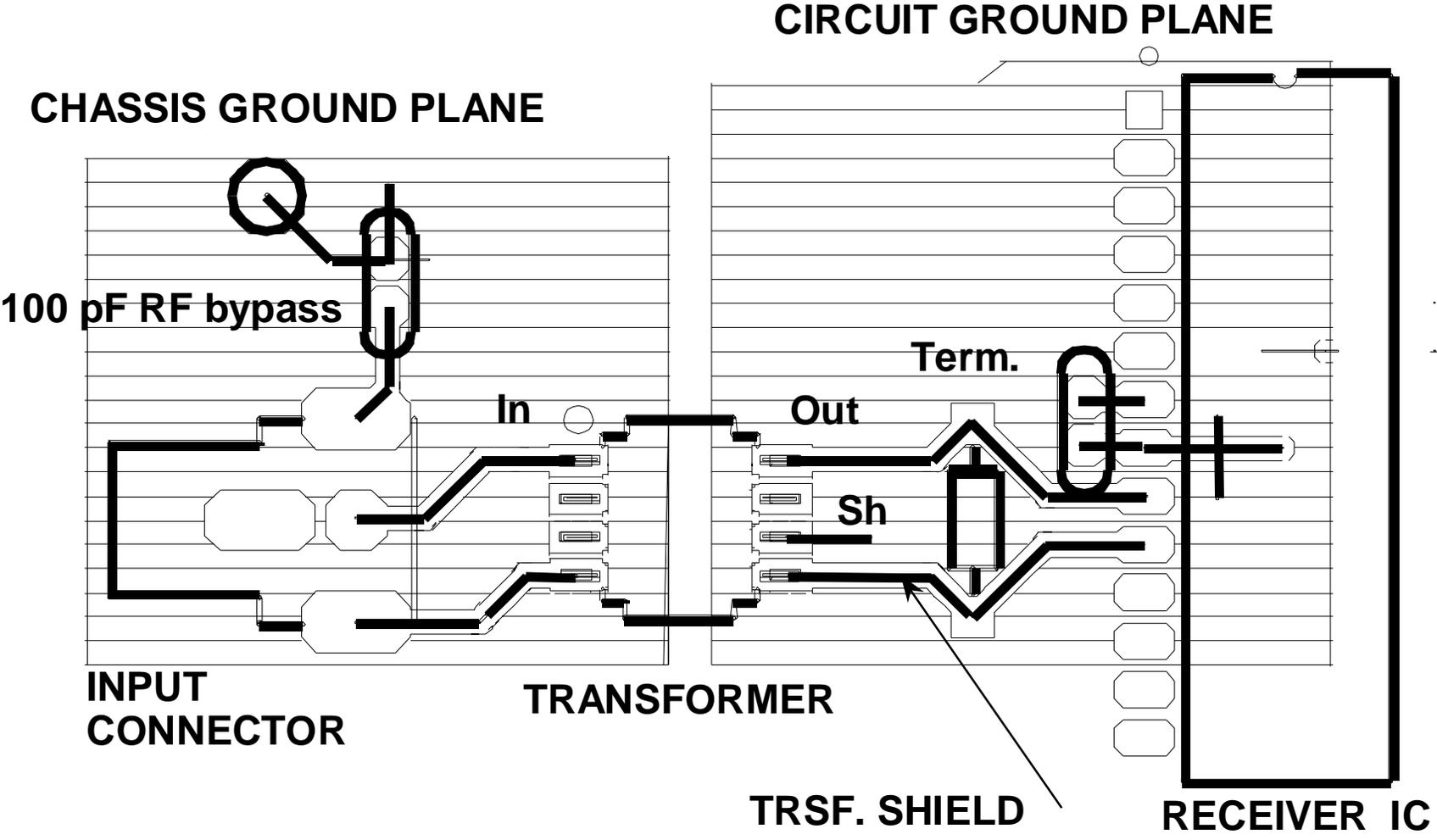


Fig. 22 Comparison of Transformers

Parameter	Unit	A	B	C	V	D	E	SC916-01	SC937-02	SC979-03
Primary Inductance	μH	2500	250	2000	600 - 2200	3500	900	5000	850	300
Capacitance P - S	pF	25	8	28	5.5	23	16	15	3.0	2.0
Shielded Cap. (NA= none)	pF	NA	NA	NA	1.5	NA	NA	3.0	0.5	NA
Flux Capacity	μVs	20	47	20	80	180	24	680	105	35
Adv. LF cutoff F _{LOW}	kHz	100.0	85.0	100.0	NA	5.0	20.0	3.5	20.0	32.0
Calc. LF cutoff F _{LOW}	kHz	7.0	70.1	8.8	7.5 - 30	5.0	19.5	3.5	20.6	35.0
Meas. LF cutoff F _{LOW}	kHz	3.0	26.0	3.0	24.3	20.0	6.5	1.5	13.0	32.0
HF Cutoff F _{HIGH}	MHz	55	50	55	9	28	150	65	100	200
RMS Jitter 11.2896MHz/2.25V	ps	1950	1260	1900	1060	2400	1320	650	502	443
CMRR	dB	23	32.5	21.5	38	23.5	27	41	50.5	52.0
Pulse Aberration	%	11	32	20	51	4	13	5	4	4
Size W x L x H	mm	12 x 15 x 8.5	9 x16 x 6	9 x 12.5 x 6	9.9 x 12 5.6	19 x 22 x 16	7.5 x 9 x 5	13 x 13 x 13	9.5 x 11 5	9.5 x 11 5
Relative Cost for 1K	ratio	1.00	1.36	1.53	1.55	2.40	3.00	3.10	3.10	3.10
Leadtime	week	12-16	5	12-16	8	6-10	4-8	1	1	1

Fig. 23 Suggested PCB Layout for Transformer Input Circuit



Summary and Conclusions

- Commercial transformers exhibit tremendous differences in parameters and performance
- Digital audio transformers substantially improve common mode noise rejection and EMI emission
- Second order effects such as pulse aberration and saturation greatly influence waveform fidelity.
- AES/EBU receivers exhibit jitter that is function of the transformer CMRR and capacitance
- Professional, broadcast and high resolution applications need maximum CMRR to minimize recovered clock jitter in the presence of noise.
- High quality transformers offer cost effective improvements in product design.

Recommendations

- Use transformers with lowest possible Primary-Secondary capacitances and leakage inductances
- For best CMRR use a low-capacitance transformer, shields and careful PCB layout
- Transformer Low Frequency bandwidth depends on minimum Frame Sync frequency. Use $1/5 - 1/20$ x AES/EBU min. spec of 100 kHz
- Transformer HF bandwidth should be 5 - 20 x of AES/EBU spec of 8MHz - depending on max. FS
- Minimize pulse aberration: under 10%